VOLUME IV

OF

FINAL REPORT NUMBER 5512001

DESIGN CONSIDERATIONS AND COMPONENT SELECTION CRITERIA

UPDATED BY M. GANTSWEG

DECEMBER 1967

IN REFERENCE TO:

MULTI-ALTITUDE TRANSPONDER
CONTRACT NO. DA-44-009-AMC-1246(X)
MATS TOP DRAWING NUMBER J5506800

FOR

U.S. ARMY ENGINEER
TOPOGRAPHIC LABORATORIES
SURVEYING SYSTEMS DIVISION
FORT BELVOIR, VIRGINIA 22060

REPORT PERIOD JUNE 1967 TO DECEMBER 1967

ITT AEROSPACE 15151 BLEDSOE STREET, SAN FERNANDO, CALIFORNIA

ITT

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FINAL REPORT NUMBER 5512001

<u>DESIGN CONSIDERATIONS AND</u>

<u>COMPONENT SELECTION CRITERIA</u>

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SECTION I

DESIGN CONSIDERATIONS

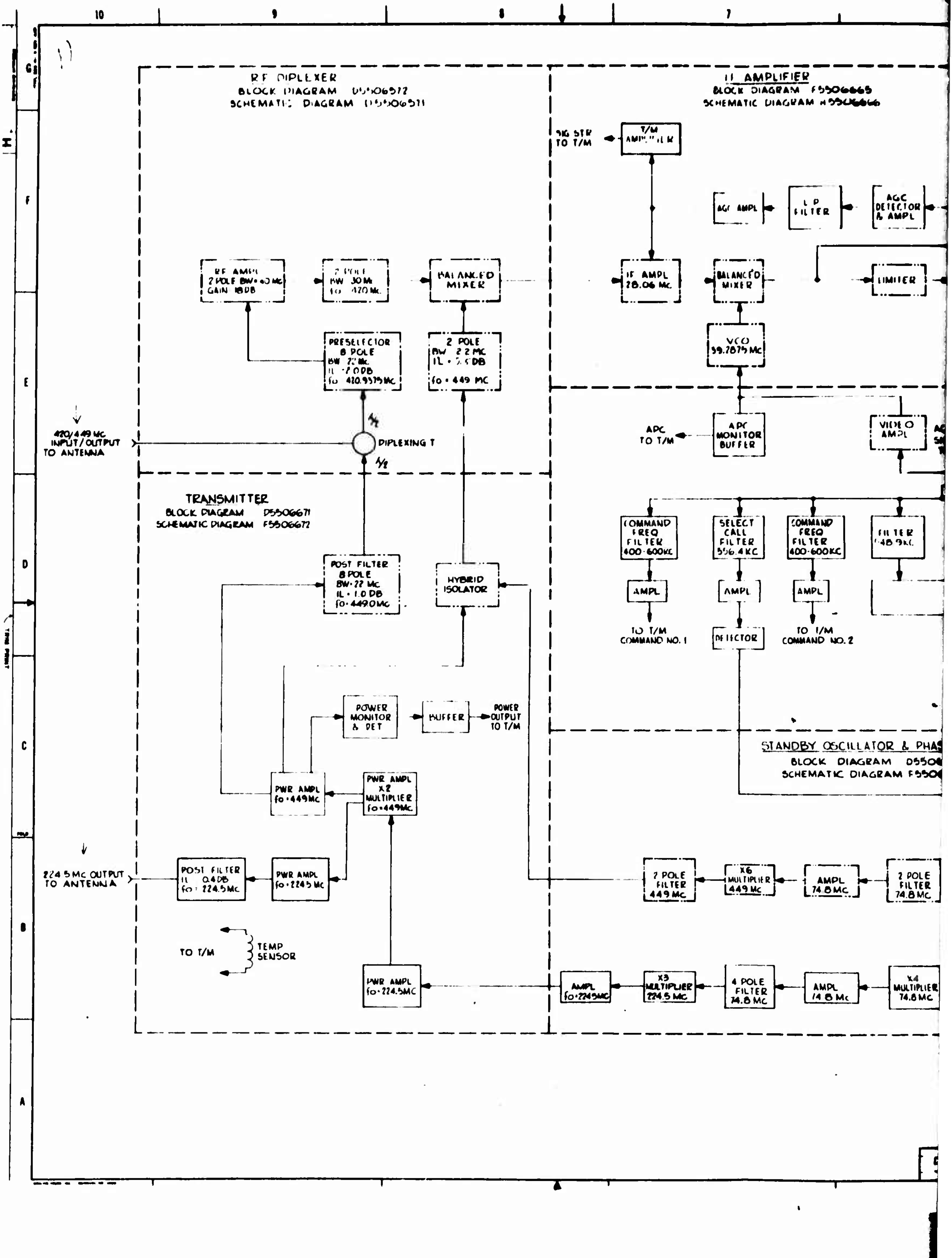
1.0 FUNCTIONAL DESCRIPTION OF TRANSPONDER

The MAT transponder utilizes a crystal-controlled double-conversion phase locked (correlation) receiver with a subcarrier phase-following loop around the complete transponder, for the purpose of phase stability and the realization of the required transponder sensitivity. Thus high sensitivity is realized in the presence of high modulation index subcarriers. Coherent AGC is employed in the receiver. The transmitter is crystal-controlled and phase-modulated, and employs transistors as active stages throughout, as does the transponder receiver. The data amplifier, which improved the transponder signal-to-noise performance by crystal filters, is a fully micro-electronic design, as is much of the transponder receiver circuitry. Cavity filters are employed to realize a high-quality, low-leakage diplexer. The power supply includes integral voltage regulators and dc-to-dc converters of the pulse width modulation type.

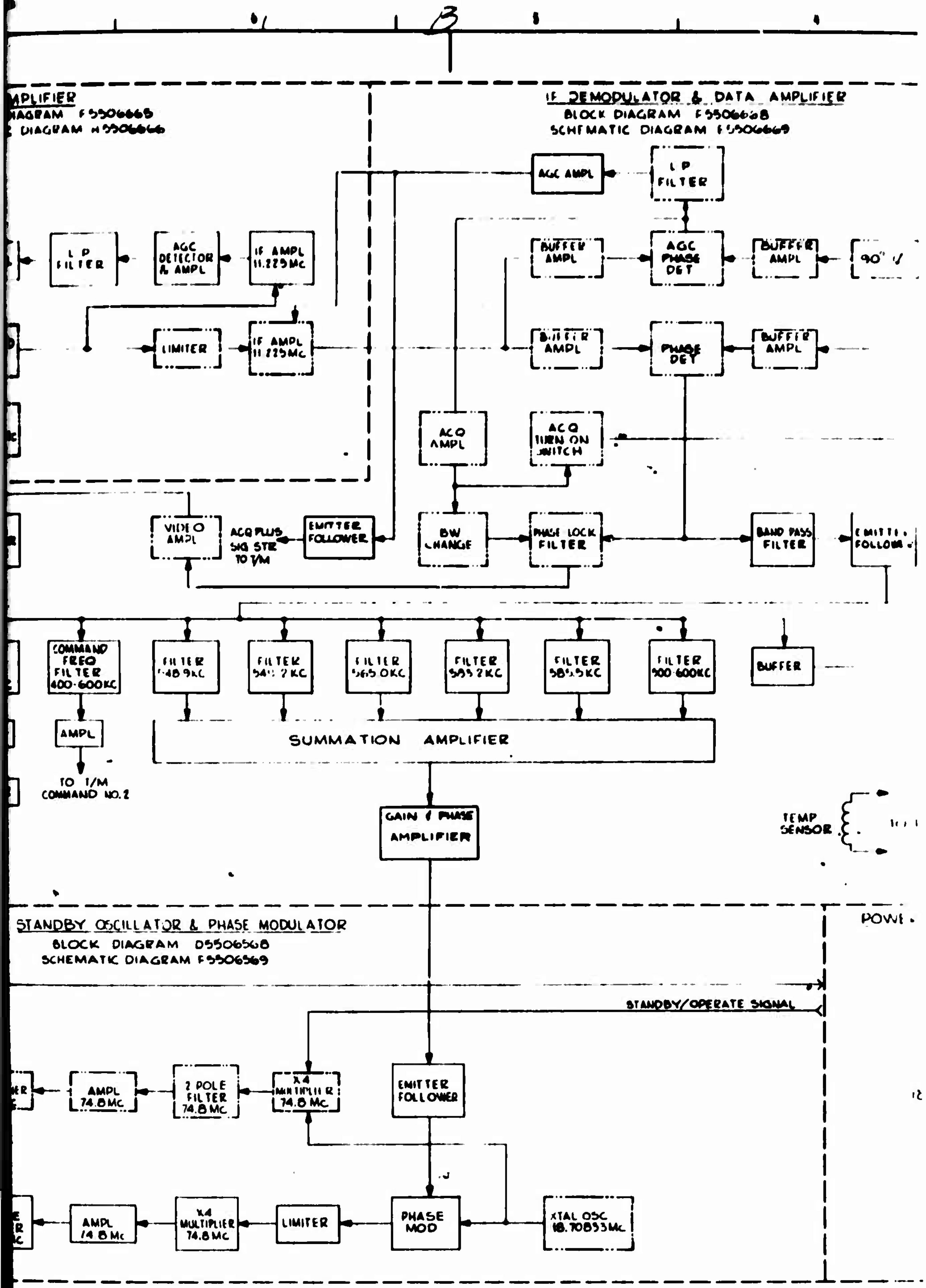
The transponder consists of six basic modules: (See Figures 1 and 2)

1. RF Diplexer (Dwg. D/5506571-B)
2. IF Amplifier (Dwg. H/5506666-C)
3. IF Demodulator and Data Amplifier (Dwg. J/5506669-B)
4. Standby LO Oscillator and Phase Modulator (Dwg. F/5506569-B)
5. Transmitter (Dwg. F/5506672-C)
6. Power Supply (Dwg. D/5008482+)

A brief description of the circuit complement of each of these modules is given below, along with the functional description of each. Refer to Figure 1 for functional block diagram.



RF DIPLEXER
BLOCK DIAGRAM D5506572
SCHEMATIC DIAGRAM D5506571
IF AMPLIFIER
BLOCK DIAGRAM F5506665
SCHEMATIC DIAGRAM H5506666
SIG STR TO T/M
T/M AMPLIFIER
AGC AMPL
L P FILTER
AGC DETECTOR & AMPL
RF AMPL 2 POLE BW=40 MC GAIN 18DB
2 POLE BW 30 MC fo 420 MC
BALANCED MIXER
IF AMPL 28.06 MC
BALANCED MIXER
LIMITER
PRESELECTOR 8 POLE BW 22 MC IL 2.0DB fo 420.9375MC
2 POLE BW 2.2 MC IL 5.5 DB fo 449 MC
VCO 59.7875 MC
420/449 MC INPUT/OUTPUT TO ANTENNA
DIPLEXING T
APC TO T/M
APC MONITOR BUFFER
VIDEO AMPL
TRANSMITTER
BLOCK DIAGRAM D5506671
SCHEMATIC DIAGRAM F5506672
COMMAND FREQ FILTER 400-600KC
SELECT CALL FILTER 556.4KC
COMMAND FREQ FILTER 400-600KC
FILTER 48.9KC
POST FILTER 8 POLE BW=22 MC IL=1.0 DB fo=449.0MC
HYBRID ISOLATOR
AMPL
AMPL
AMPL
TO T/M COMMAND NO. 1
DETECTOR
TO T/M COMMAND NO. 2
POWER MONITOR & DET
BUFFER
POWER OUTPUT TO T/M
STANDBY OSCILLATOR & PHAS
BLOCK DIAGRAM D550
SCHEMATIC DIAGRAM F550
PWR AMPL fo=449MC
PWR AMPL X2 MULTIPLIER fo=449MC
224.5 MC OUTPUT TO ANTENNA
POST FILTER IL 0.4DB fo = 224.5MC
PWR AMPL fo=224.5 MC
2 POLE FILTER 449 MC
X6 MULTIPLIER 449 MC
AMPL 74.8 MC
2 POLE FILTER 74.8 MC
TO T/M
TEMP SENSOR
PWR AMPL fo=224.5MC
AMPL fo=224.5MC
X3 MULTIPLIER 224.5 MC
4 POLE FILTER 74.8 MC
AMPL 74.8 MC
X4 MULTIPLIER 74.8 MC



IF DEMODULATOR & DATA AMPLIFIER
BLOCK DIAGRAM F5506668
SCHEMATIC DIAGRAM F5506669
AGC AMPL
L P FILTER
BUFFER AMPL
AGC PHASE DET
90°
L P FILTER
AGC DETECTOR & AMPL
IF AMPL 11.225Mc
LIMITER
PHASE DET
ACQ AMPL
ACQ TURN ON SWITCH
VIDEO AMPL
ACQ PLUS SIG STR TO T/M
EMITTER FOLLOWER
BW CHANGE
PHASE LOCK FILTER
BAND PASS FILTER
COMMAND FREQ FILTER 400-600KC
AMPL
TO T/M COMMAND NO. 2
SUMMATION AMPLIFIER
BUFFER
GAIN & PHASE AMPLIFIER
TEMP SENSOR
STANDBY OSCILLATOR & PHASE MODULATOR
BLOCK DIAGRAM D5506568
SCHEMATIC DIAGRAM F5506569
STANDBY/OPERATE SIGNAL
AMPL 74.8Mc
2 POLE FILTER 74.8Mc
X4 MULTIPLIER 74.8 Mc
EMITTER FOLLOWER
PHASE MOD
LIMITER
XTAL OSC 18.70833Mc

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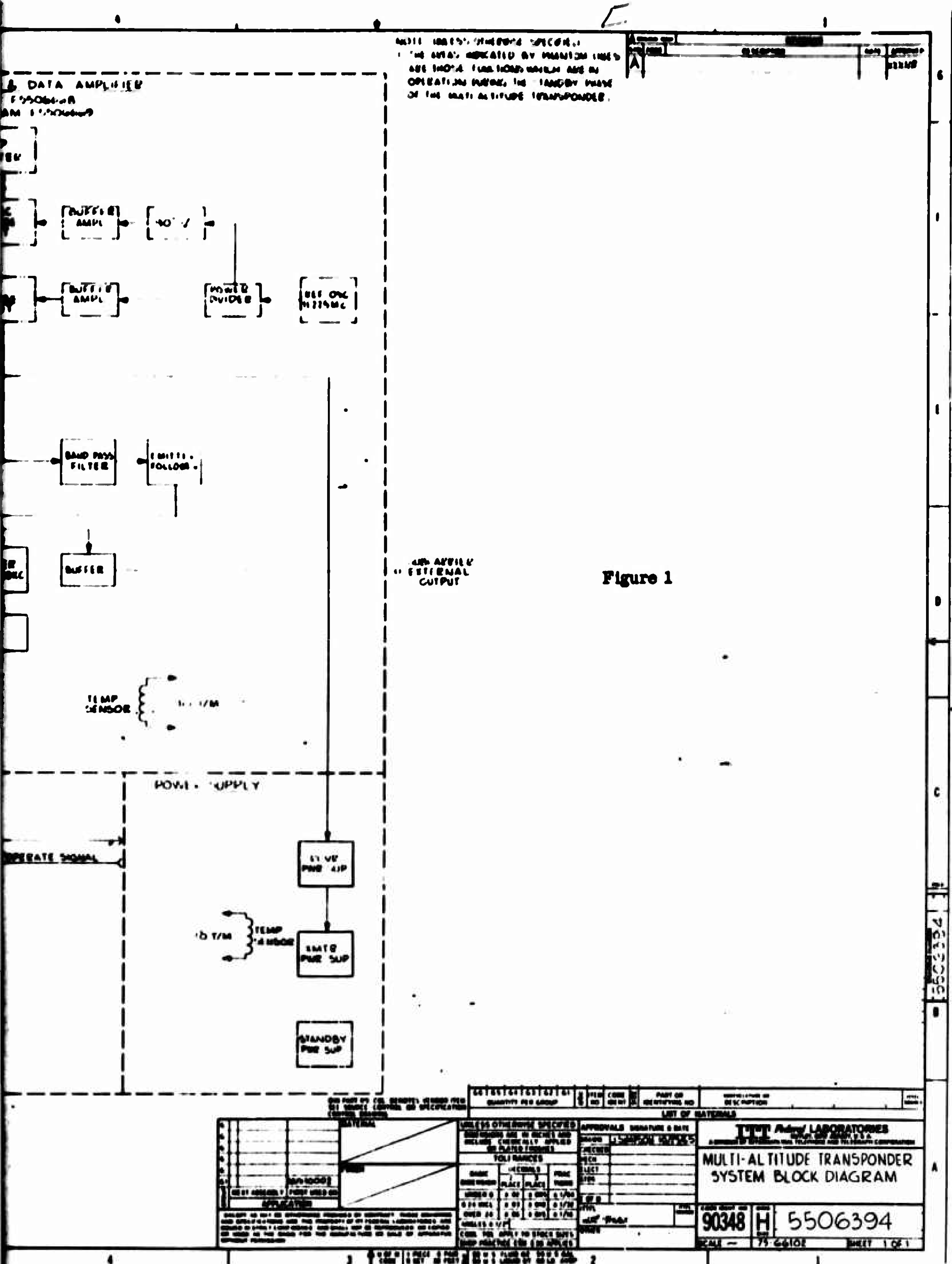

DATA AMPLIFIER
BUFFER AMPL
BUFFER AMPL
POWER DIVIDER
REF OSC
BAND PASS FILTER
EMITTER FOLLOWER
BUFFER
TEMP SENSOR
POWER SUPPLY
OPERATE SIGNAL
TEMP SENSOR
XMTR PWR SUP
STANDBY PWR SUP
EXTERNAL OUTPUT
LIST OF MATERIALS
APPLICATION
ITT Federal LABORATORIES
MULTI-ALTITUDE TRANSPONDER SYSTEM BLOCK DIAGRAM
90348
H
5506394
5506394

Figure 1

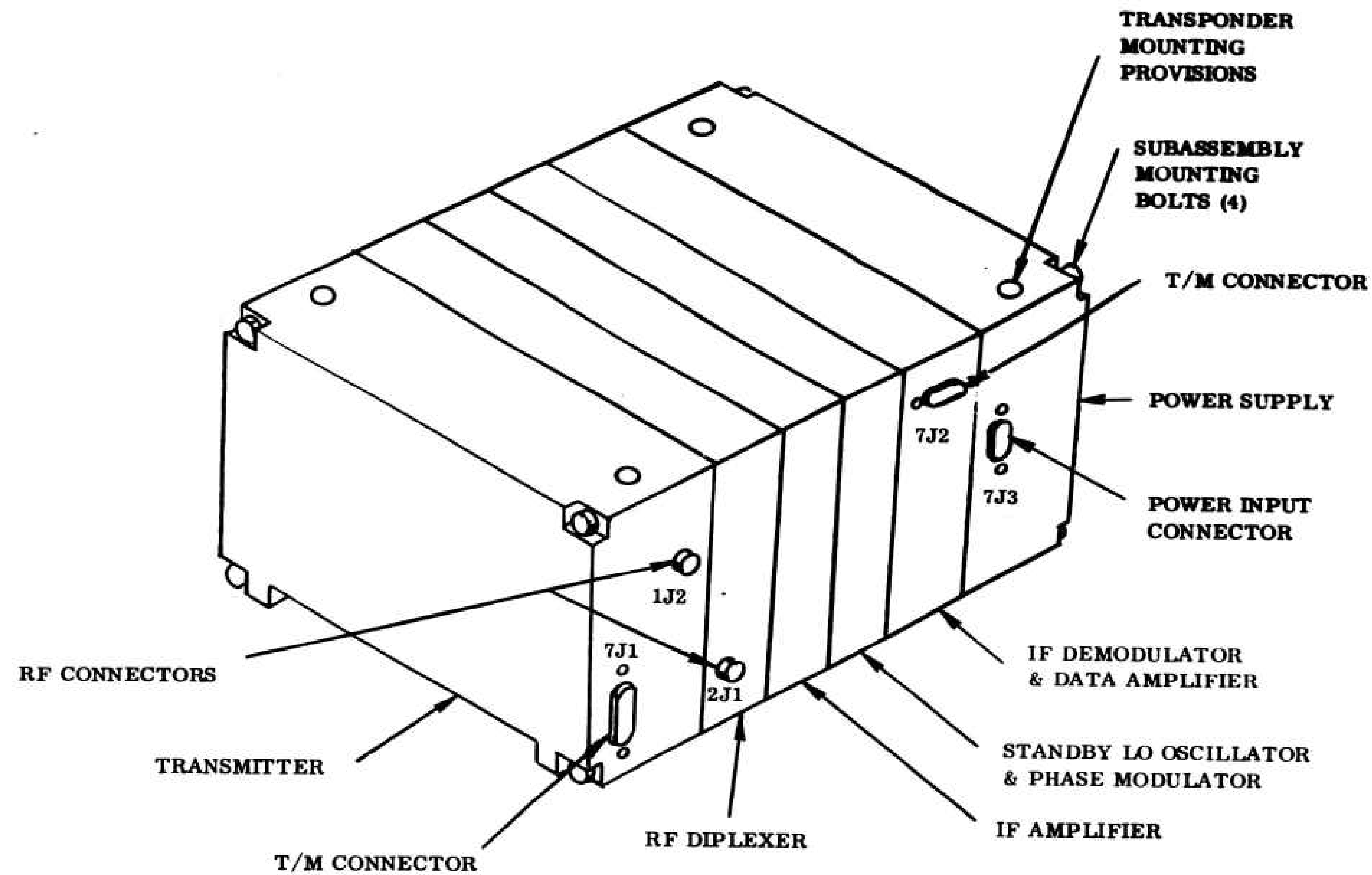


Figure 2. Module and Connector Configuration

The RF Diplexer module contains the Diplexing T, the preselector portion of the diplexer, a two-stage RF amplifier, another 2-pole preselector, a balanced mixer, a 2-pole L.O. filter, a standby duty cycle circuit, and a power supply delay circuit.

A low level 420.9375 MHz signal (-45 to -115 dbm) is received by the transponder and fed into the Diplexing T and through an 8-pole 20 MHz bandwidth preselector to the RF amplifier. Simultaneously, a 449 MHz high level signal (+37 dbm) is being transmitted through the T to the receiver shared antenna. The 8-pole preselector guarantees that the level of the 449 MHz signal to the RF amplifier is less than -23 dbm, thereby negating problems which otherwise would have been encountered if the area of operation for the RF amplifier was nonlinear. The RF amplifier is a low noise wideband circuit which provides +17 db of gain to the 421 MHz signal and only +2 db to the undesirable 449 MHz signal. To further reduce undersirable effects from the 449 MHz signal, an additional 2-pole bandpass filter (30 MHz bandwidth) is used to pass 421 MHz and reject 449 MHz by 27 db. The output of this 2-pole filter is fed to the RF input side of a balanced hot-carrier diode mixer. The L.O. input to the mixer is obtained from the transmitter module. Since this mixer is used as the Phase Following Subtractor to close the transponder phase following loop, the L.O. input is wideband and contains the spread spectrum signal. A 2-pole bandpass filter (30 MHz BW) is used to pass the wideband signal about 499 MHz while rejecting out-of-band frequency components sufficiently to provide a clean L.O. injection to the mixer. The mixer output is the product of the 449 MHz sirnal and the 421 MHz signal giving the sum and difference frequencies along with higher order products, each with their own modulation spectrums. The desired signal is the difference frequency (28.06 MHz) and the associated modulation spectrum. Note that the output modulation spectrum about 28.06 MHz is smaller than those at 421 MHz or 449 MHz due to phase subtraction of the spectral components by the mixer.

The duty cycle circuit is used to adjust the transponder standby power duty cycle in order to conserve total consumed energy. The power supply delay circuit is used to delay the PFFB loop operation until all voltages are steady state.

The IF amplifier module consists of four synchronously tuned 28.06 MHz IF amplifier stages, a second mixer, an amplitude limiter stage at 11.225 MHz, a single stage coherent AGC'd IF amplifier at 11.225 MHz, a noncoherent AGC loop, an AGC telemetry amplifier, and a voltage controlled oscillator (VCO).

The minimum 28.06 MHz IF input signal level is -110 dbm while the input noise level is -95 dbm. The quiescent gain of the IF is 69 db providing an input signal level to the second mixer of -42 dbm plus noise at -28 dbm in a 5 MHz bandwidth. Each stage of the IF amplifier has a nominal gain of approximately 16 db and a bandwidth of 14 MHz. It is reverse AGC'd by the noncoherent AGC loop to provide +20 to -50 db of automatic gain control in the "receive" and "transmit" mode.[1] Standby mode operation provides a fixed level gain with symmetric limiting in each amplifier stage.

The second mixer is an active double-balanced type which provides a +5 db power gain at 11.225 MHz and -20 db attenuation of the 39 MHz L.O. (-5 dbm input level), and 28.06 MHz IF. The 39.28 MHz VCO feeds this mixer. Its frequency control is derived from the carrier phase lock loop in the demodulator module.

The output of the second mixer is fed to the 11.225 MHz limiter stage and the noncoherent AGC input amplifier. The symmetric limiter prevents the stage output from providing greater than -3 dbm output independent of the stage signal or normal noise input level.

[1] "Standby" Mode — Ready to receive a coherent carrier
"Receive" Mode — Coherent carrier received and ready for commands
"Transmit" Mode — All circuitry operating

The limiter output is fed to a linear 11.225 MHz amplifier which provides 15 db of coherent AGC control. The AGC is of the forward type to ensure linear operation over the entire dynamic range of operation. The 11.225 MHz coherent signal output is a constant -6 dbm, independent of normal signal or noise input levels.

The noncoherent AGC circuitry consists of 2 stages of 11.225 MHz IF, providing 32 db of power gain and a bandwidth of 2.0 MHz. Its output is fed to a detector amplifier which provides a DC output voltage proportional to the input signal level into the receiver. The output voltage is used to control a current amplifier which provides reverse AGC control for the 28.06 MHz IF stages.

The IF Demodulator and Data Amplifier module contains (1) part of the phase lock loop consisting of a phase detector, reference oscillator, phase lock, a video amplifier, and associated buffer stages ; (2) part of a correlation detector loop consisting of a phase detector, a 90° phase shifter, lowpass filter, coherent AGC amplifier, acquisition amplifier, and associated buffer stages; (3) the data amplifiers consisting of ranging and timing crystal filter networks, summation amplifiers, command and select call filters and amplifiers, a select call AM detector, and; (4) a temperature sensor and telemetry amplifiers.

The 11.225 MHz signal from the IF module is power split between the correlation and phase lock loops. The phase and correlation detector reference signals are obtained from the 11.225 MHz reference oscillator with the latter a phase quadrature resultant from the 90° phase shifter. Under locked conditions, the output of the phase detector contains the demodulated subcarriers and a DC voltage corresponding to the frequency difference between the 11.225 MHz reference and the 11.225 MHz signal before lockup. The DC voltage is fed to the

phase lock filter, which determines primarily the PLL[2] acquisition and signal-to-noise ratio performance, and a video amplifier which determines primarily the steady state phase error. The video amplifier output feeds the VCO located in the IF chassis, completing the PLL. The demodulated ranging subcarriers are fed to filter networks (3 db BW = 100 cps) for S/N ratio improvement. The output of each network is summed together, amplified and fed to the Standby LO Oscillator and Phase Modulator module. The command and select call subcarriers are fed to their respective crystal filters, again for the purpose of S/N improvement. The command outputs are amplified and fed to the transponder telemetry output connector. The select call output is amplified and detected by an AM diode detector and fed to the power supply module.

The correlation detector output is a DC voltage proportional to the received input signal. Its output is filtered, amplified and fed to the IF module for coherent AGC and to the power supply module to command the transponder from the "standby" to the "receive" mode.

The Standby LO Oscillator and Phase module provides a phase modulated transmitter driving signal and the receiver standby first local oscillator signal. The modulation is the series of ranging tones from the data amplifier. The RF signal is developed from a temperature compensated crystal oscillator.

The temperature compensated cyrstal oscillator supplies a one milliwatt signal at 18.7 MHz for both the transmitter multiplier chain and the receiver first local oscillator multiplier chain. The power split off to the transmit leg is phase modulated by the data summation amplifier output. This modulation signal is composed of up to six tones in the frequency range 400 to 600 KHz at levels of about 0.3 volts rms each. Peak modulation levels will deviate the carrier approximately 0.625 radians. The phase modulator output is amplitude limited in the following integrated circuit stage by cutoff and saturation limiting.

[2] PLL: Phase Lock Loop

Limiting is used to remove incidental AM from the phase modulated signal. The spectrum is frequency multiplied by four to approximately 75 MHz and is then amplified to 50 milliwatts. A 4-pole filter follows the amplifier to suppress unwanted harmonics of the 18.7 MHz oscillator frequency by at least 60 db. The filter has a 3 db bandwidth of 8 MHz. A frequency tripler follows the filter to raise the spectrum to 224.5 MHz. The following amplifier provides sufficient power gain to supply 50 milliwatts of output power. This 224.5 MHz output is filtered by a 2-pole cavity (Butterworth response, 15 MC BW) to reduce the 18 MHz and 75 MHz sidebands. A pad is used to adjust the drive for the transmitter module.

The other oscillator output at 18.7 MHz is multiplied by four to 75 MHz. It is then filtered in a 2-pole, 6 MHz wide filter to suppress unwanted oscillator frequency harmonics. The following amplifier raises the level of the 75 MHz to about 15 milliwatts to drive the step recovery diode multiplier to supply a 1 milliwatt output at 449 MHz. The diode output is filtered by a 2-pole 8 MHz wide filter to suppress unwanted 75 MHz harmonics. The filtered 1 milliwatt output at 449 MHz is used as the receiver standby first local oscillator.

The Transmitter module contains a 224 MHz medium power amplifier, X2 multiplier, a 224 MHz high power amplifier, 449 MHz high power amplifier, power monitor, hybrid isolator, a 449 MHz post filter, and a 224.5 MHz post filter.

The transmitter module receives an input signal at 224.5 MHz at a power level of approximately 20 milliwatts. This signal is amplified to 1 watt, fed to the 224.5 MHz final amplifier, whose output of 5 watts is fed to a 2-pole bandpass post filter providing approximately 4.5 watts output into a 50 ohm load. The 3 db bandwidth of the 224.5 MHz output is 12 MHz. The 224.5 MHz 1 watt signal is derived from a frequency doubler whose 449 MHz output is 1.5 watts. This output feeds the 449 MHz final stage which provides 5.5 watts to the 8-pole post filter. The post filter 1.0 db insertion loss reduces the transponder output power to

4.5 watts into a 50 ohm load. The power monitor is an AM detector using an RF diode into a buffer amplifier whose output is fed to the telemetry connector. The hybrid is used to isolate the Standby L.O. circuitry from the Operate L.O. circuitry.

The Power Supply module provides transponder power for "standby", "receive" and "transmit" modes. Incorporated in the module are time delay and logic circuits which control standby and transmit power. The outputs are regulated against input and load changes. They are isolated from input power leads and the module chassis. Momentary short circuit and reverse polarity input protection is provided.

One of the prime considerations for the technical approach to the MATS Power Supply was the conservation of battery power. In order to achieve this, the unit was designed with high efficiency as one of the important aspects. For this reason, pulse width modulators are used to convert the available fluctuating input voltage into a highly regulated DC voltage which powers a DC to DC converter. The PWM utilizes the transistor in the switching mode only. During the on-cycle of the switch, energy flows through an inductor into the output capacitor. Simultaneously, energy is stored in the inductor. During the off-cycle, the inductor discharges its stored energy into the output capacitor. By varying the duty cycle, regulation is achieved at a high efficiency level. The DC to DC converter (often called a square-wave inverter) provides three distinct advantages:

- The converter allows step-up or step-down to any desired output voltage.
- The output wave shape is a square-wave which after rectification requires very little filtering.
- The transistors operate in a switching mode only and therefore cause the lowest losses, resulting in a high efficient conversion.

Sensing of the output voltage and error signal amplification is accomplished by transistor differential amplifiers which are known for their good temperature stability. The differential amplifier controls a magnetic amplifier which is powered from the converter transformer. The output of the magnetic amplifier controls the duty cycle of the pulse width modulator. Sensing could have been accomplished by the magnetic amplifier only; however, then, the problem of temperature stability would have become critical.

Only those outputs that are the most critical as far as regulation are concerned are being sensed. The other outputs are taken, after rectification, directly from the converter transformer. As the input to the DC to DC converter is regulated, the output of the DC to DC converter is also regulated and is subject to transformer regulation and coupling only.

2.0 DESIGN CONSIDERATIONS

2.1 General Description

The MAT transponder was designed to be a compact, lightweight, efficient transponder for use in Satellite configurations. The transponder consists of a receiver and transmitter for accepting ranging, timing and command information from a station located on a ground complex and retransmitting the ranging and timing information on two offset carriers back to the ground complex. In addition, certain telemetry circuits are provided within the transponder in order that conditions concerning the transponder may be telemetered back to the ground station by telemetry systems external to the transponder.

The transponder's overall dimensions are 8-1/2" x 4-1/4" x 6-1/2" or 235 cubic inches (Refer to Figure 3) and its total weight is 11-1/2 lbs. In standby it requires 0.8 watts average power. In the transmit mode it delivers a nominal 4.5 watts at 224.5 MHz and a nominal 4.5 watts at 449 MHz, the latter under a diplexed condition with its receiver, while using 39 watts of primary power for an overall efficiency of 23.1 percent. Other input-output power options are available, down to 1.5 watts each channel, by a simple adjustment.

Modular construction has been employed for ease of fabrication and maintenance. Figure 4 is a photo of the complete transponder and Figures 5 through 10 are photos of each module.

The following discussion of design considerations is referenced by title directly to the purchase description for Multi-Altitude Transponder, dated 15 January 1965.

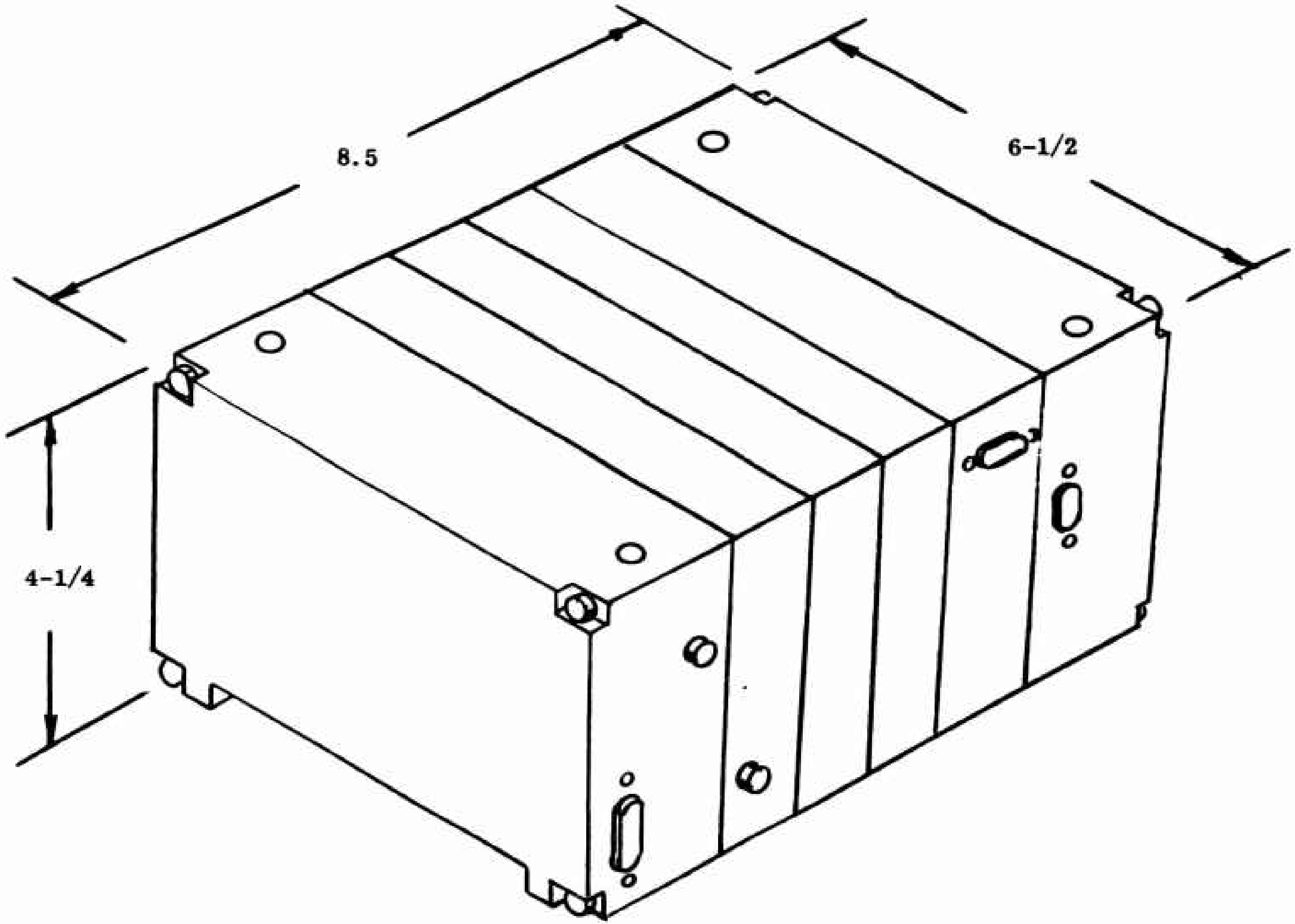


Figure 3. Transponder Dimensions

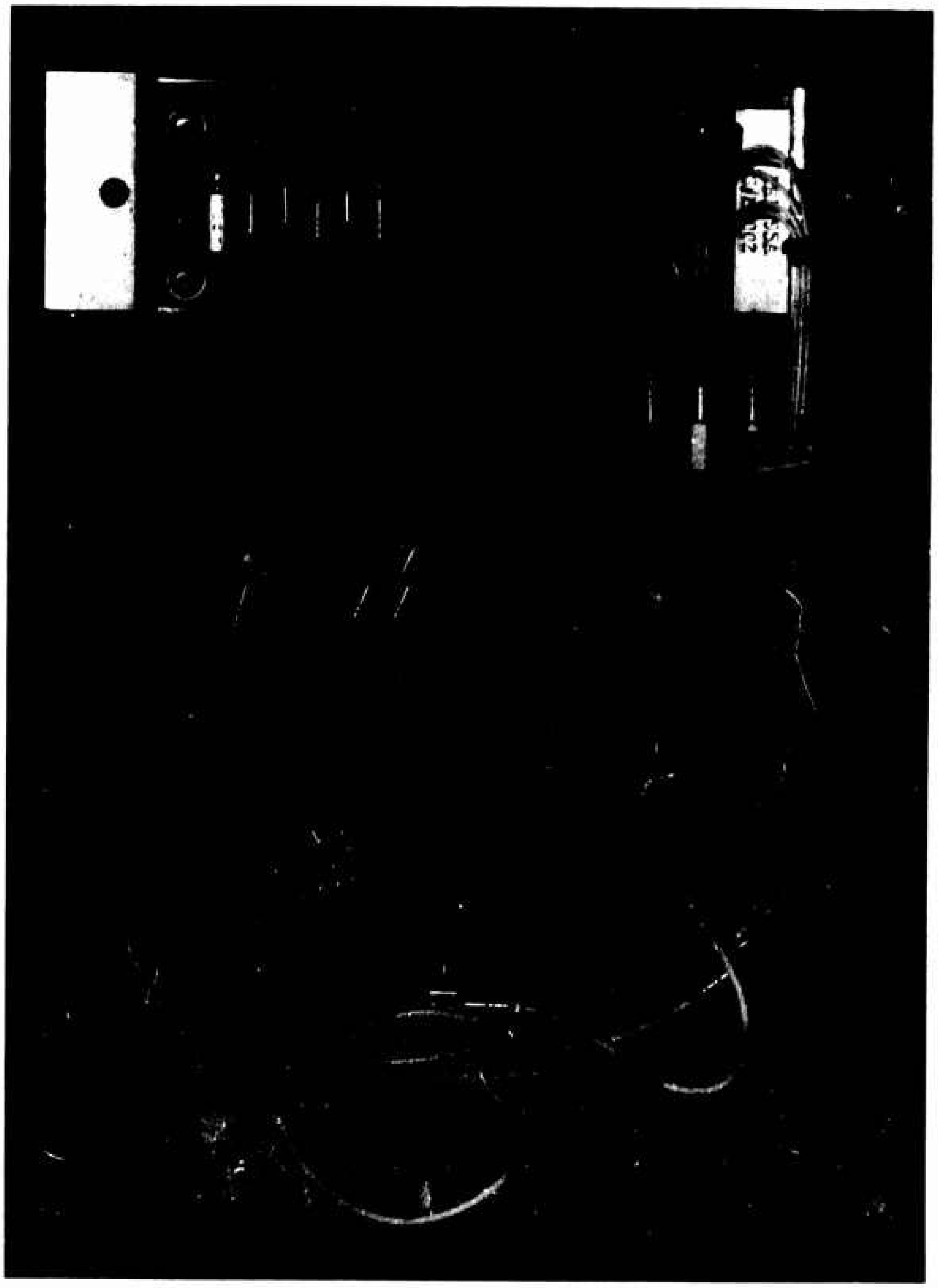


Figure 4. Transponder Assembly

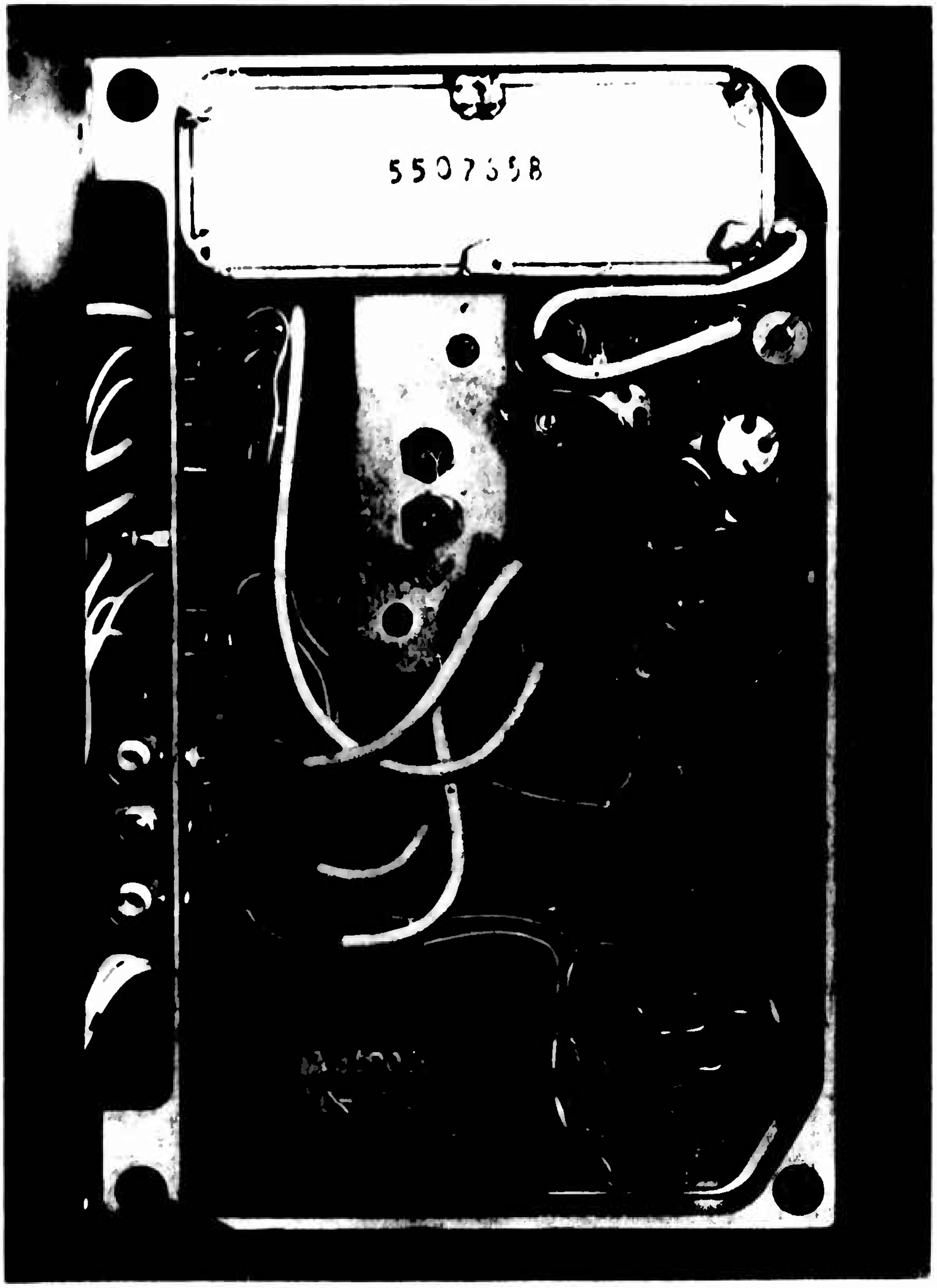


Figure 5. Standby LO Oscillator and Phase Modulator Module

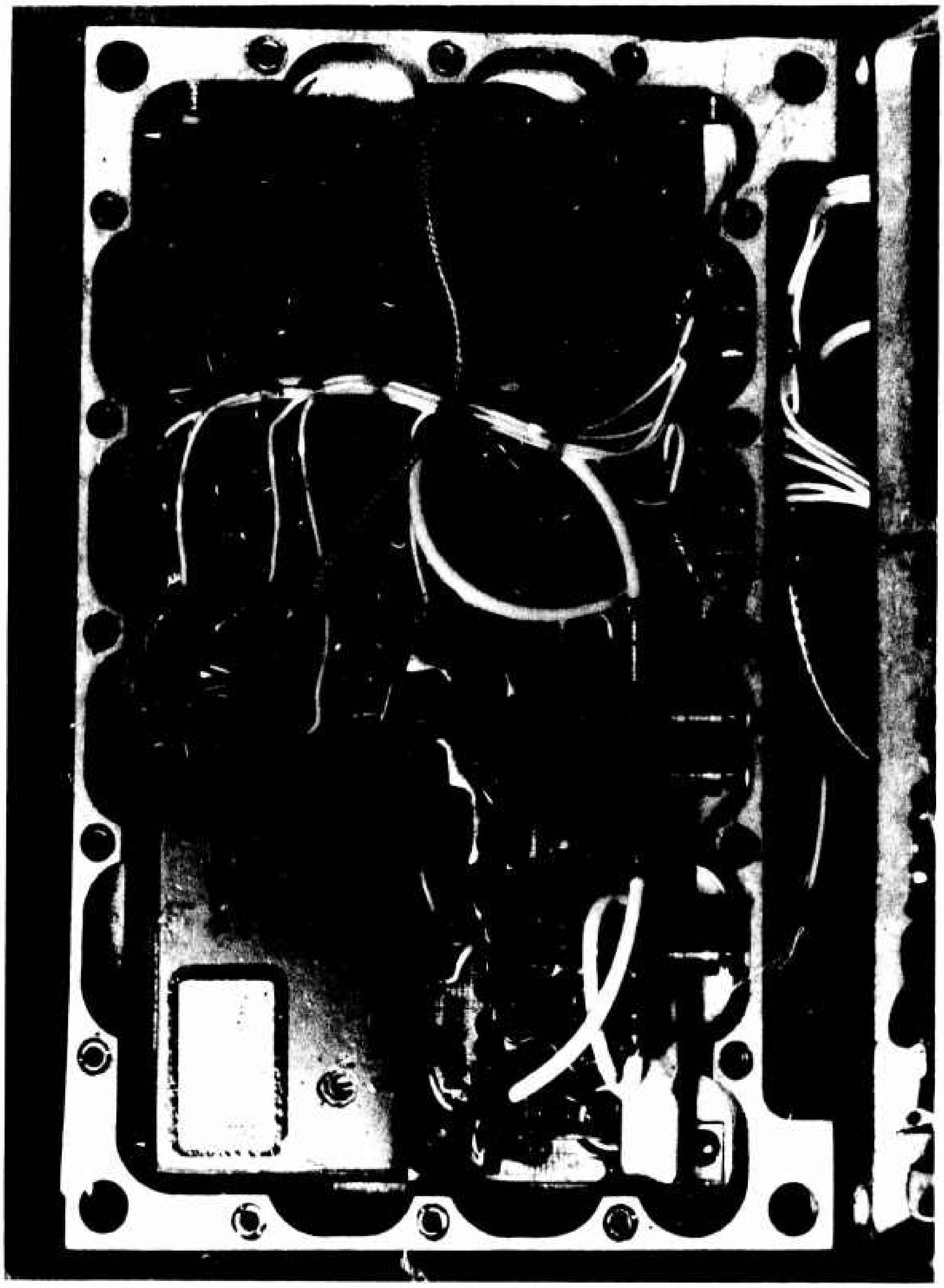
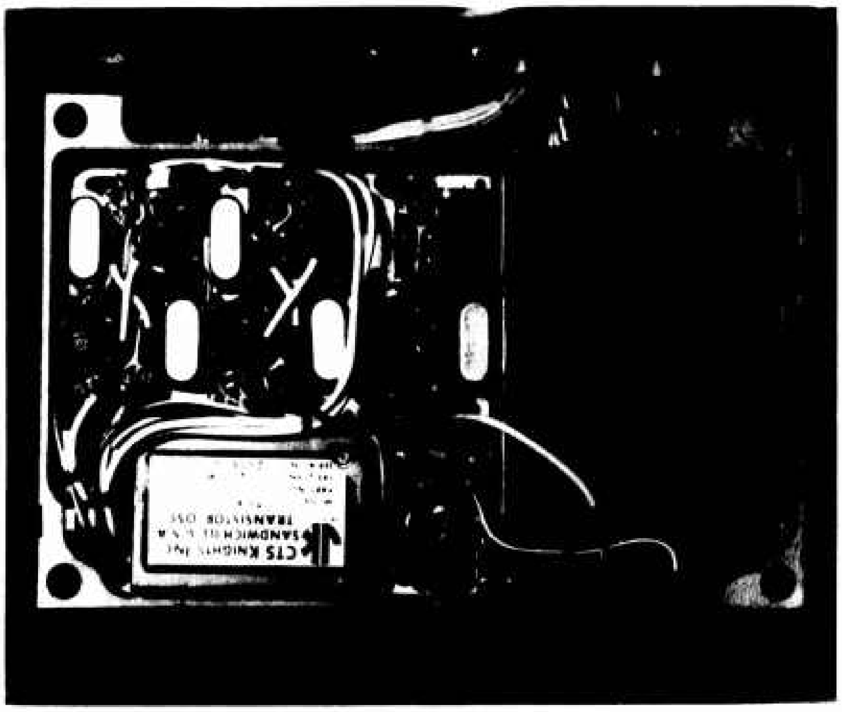


Figure 6. IF Amplifier Module



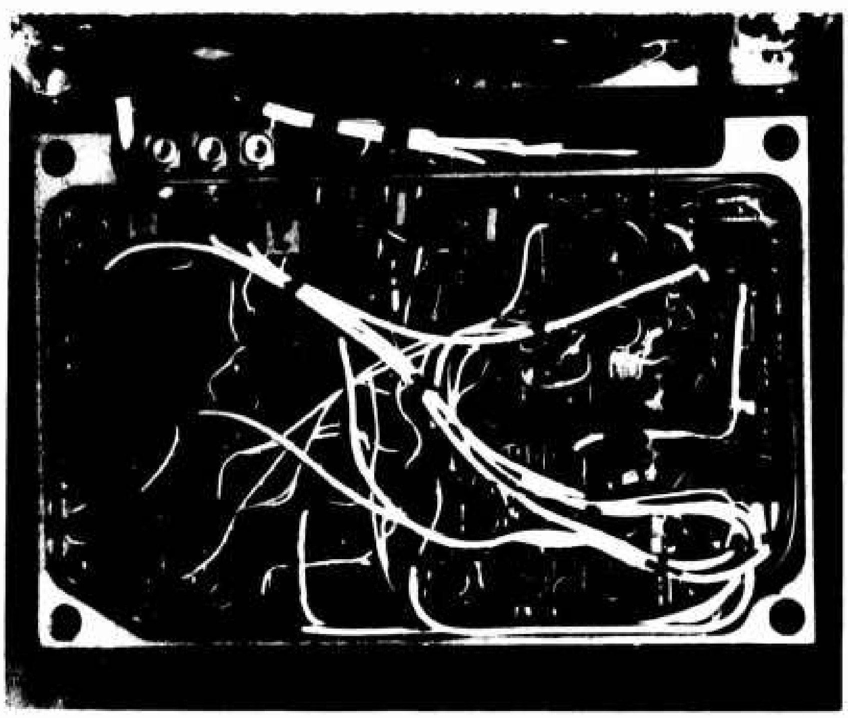


Figure 7. IF Demodulator and Data Amplifier Module

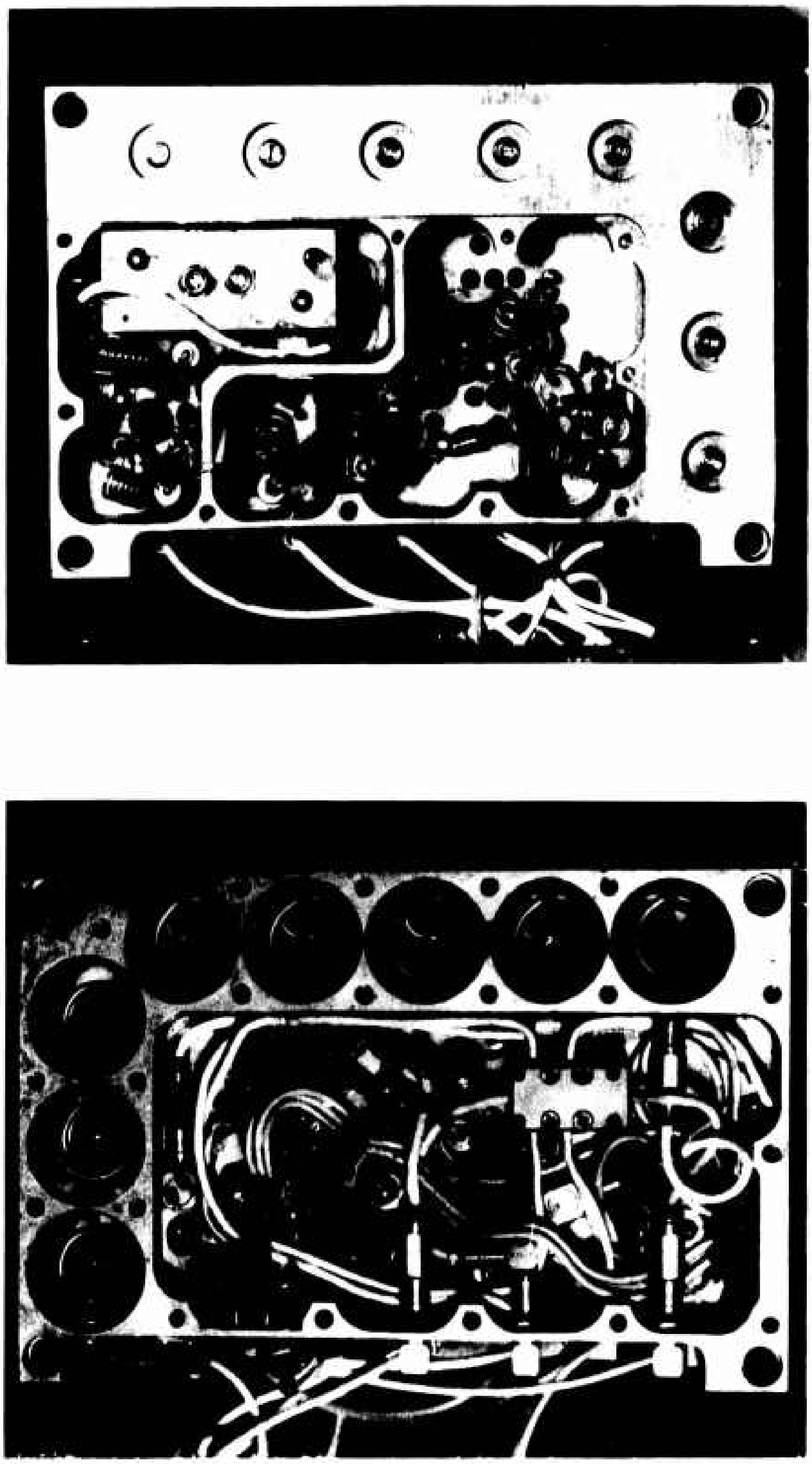
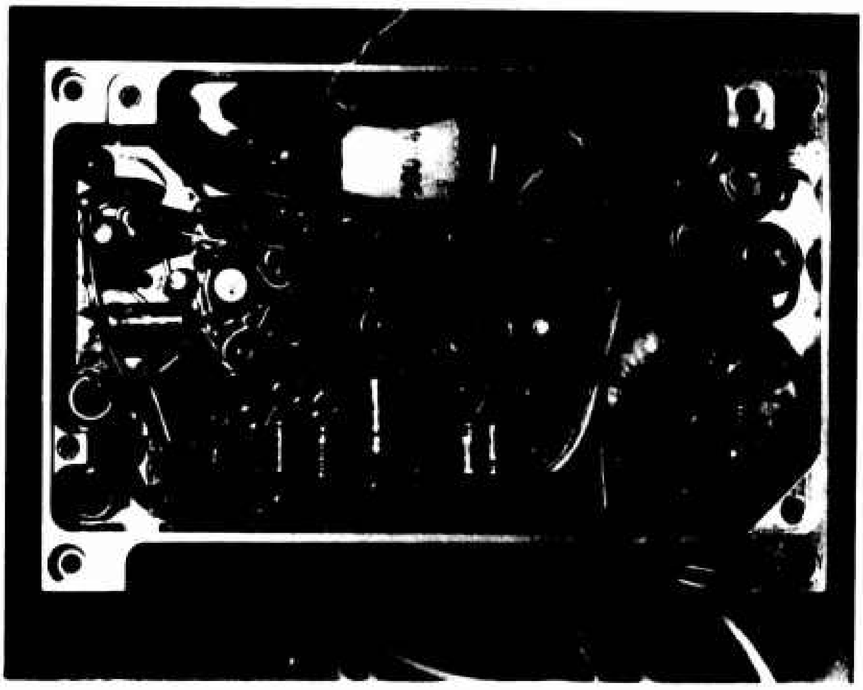


Figure 8. Transmitter Module



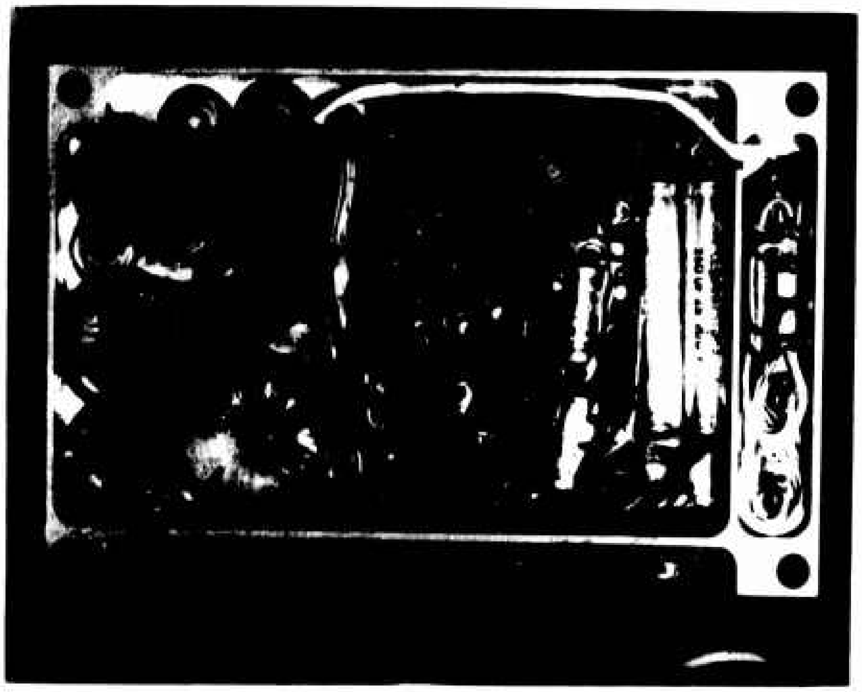


Figure 9. Power Supply Module

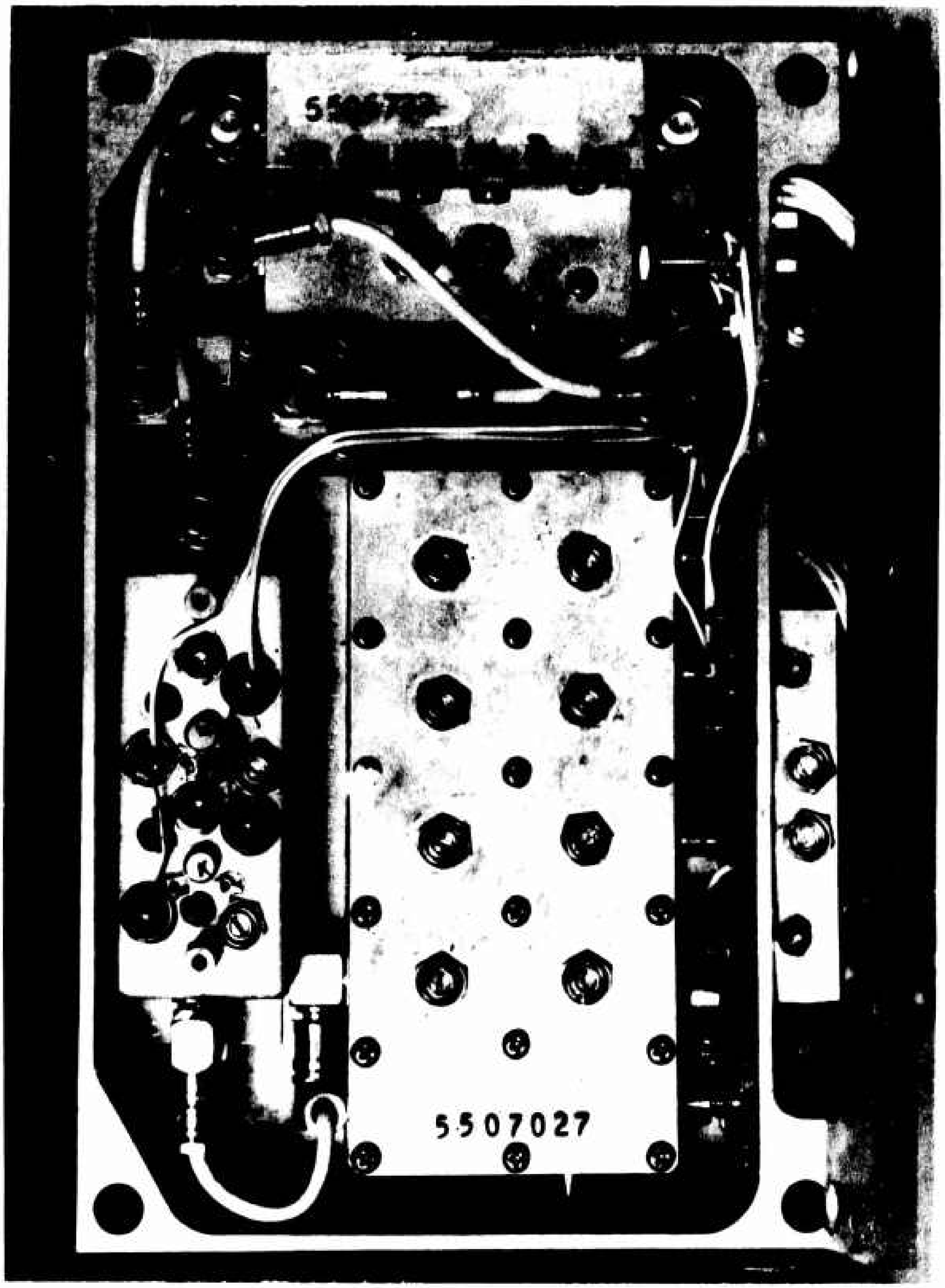


Figure 10. RF Diplexer Module

2.1.1 Composite Signal (PD para. 3.1.1)

The composite signal from the ground complex consists of a carrier whose nominal frequency is 420.9375 MHz phase modulated by any combination of fixed frequency subcarriers as follows:

	Subcarrier Frequency	Description
a.	585.533 KHz	Subcarrier referenced to ground station complex and used to measure range.
b.	583.246 KHz	Same as a.
c.	549.223 KHz	Same as a.
d.	548.937 KHz	Same as a.
e.	565.000 KHz	Subcarrier used to establish system timing.
f.	Subcarrier in the range of 500-600 KHz	Subcarrier used as a Satellite ranging function
g.	Subcarrier in the range of 500-600 Khz	Subcarrier used as a Satellite Command function
h.	Subcarrier in the range of 500-600 KHz	Same as g.
i.	Subcarrier in the range of 400.00 KHz to 600.00 KHz	Subcarrier known as "Select-call" and used to command the transponder from receive to a transmit condition.

The modulation index of each individual ranging or timing subcarrier can be as high as 2.5 radian and will normally exist within the range of 0.5 to 2.5 radians. The modulation index of the select call and command subcarriers should be in the range of 0.25 to 0.5 radians. The composite index can be any index that might result from any combination of subcarriers modulated within this range.

Of the above received subcarriers, only those shown in a. through f. are modulated on the transmitter carrier for retransmission to the ground complex.

Of particular interest is the different modulation index requirement for the select call and command subcarriers. The low 0.25 to 0.5 index is necessary since these particular subcarriers are not retransmitted by the transponder, thereby requiring their composite indexes not to exceed 1 radian. Linearity is thus preserved through the receiver phase detector.

A note on the retransmission of the command and select call subcarriers.

Since $\frac{M_I'}{M_I} = 30$ for a phase following feedback ratio = 30

where M_I' = Maximum modulation index of select call or command subcarriers

$M_I' = 0.5$ radians

M_I = Minimum modulation index of a range or timing subcarrier within the transponder phase following feedback loop

$M_I = 0.5/30 = .0166$ radians

then the baseband crystal filters used for ranging and timing subcarriers are required to reject the command and select call frequencies by 300:1 to ensure

that a maximum select call or command index of, say, 0.05 radians is modulated on the transmitter carrier. Since, for stability reasons, a single-pole crystal filter is required for each fedback subcarrier, and a 3 db bandwidth of 100 cps is reasonably specified, then, due to a PFFB[3] ratio of 30:1, a resultant closed loop bandwidth of 3 KHz and 6 db/octave filter rejection rate, one would expect no greater than a 23 db or 14:1 rejection of a subcarrier frequency 30 KHz distant.

Thus, in order to ensure that the command and select call subcarriers are minimally modulated on the transmitter carrier: (1) their subcarrier frequencies should be chosen as far from the feedback subcarriers (a-f) as possible; (2) the modulation index used for their transmission should be kept as small as possible and; (3)the modulation index of subcarriers a-f should be maximized. Thus

$$\frac{M'_I}{M_I} = \frac{.25}{2.5/30} = 3$$

Filter rejection ratio = 14:1

then the ratio of modulation index of each desired subcarrier to that of the select call or command subcarrier is 14/3 or 4.7:1.

2.1.2 Transmitter (PD para. 3.1.2)

The nominal frequency of the two (2)offset transmitter frequencies is 449.000 MHz and 224.500 MHz. The transmitter section of the transponder is designed to provide either 1.5, 3.5 or 4.5 (selectable) watts of output power at the transmitter antenna terminal for the two (2) transmission frequencies of the transmitter.

[3] PFFB: Phase Following FeedBack

The 449 MHz and 224.5 MHz frequencies are easily derived from a common TCXO at 18.7083 MHz. A not so easy task is providing the 4.5 watts output power at the 449 MHz transmitter antenna terminal. This problem area exists because of the high insertion loss exhibited by the post filter and associated diplexer loss. The high insertion loss is due to (1) the large number of poles required by the filter plus (2) the wide bandwidth required to pass the information in association with the small frequency separation between the transmitter and receiver frequencies. The choice of 8 poles as the number required by the post filter is derived in Appendix A. The primary design criteria being based upon the amount of noise generated by the transmitter in the receiver acceptance frequency band. Appendix B verifies the wide bandwidths required by the post and preselector filter to pass the modulation. Using a cavity type filter, constructed within the size limitation of the package, the calculated theoretical insertion loss for an 8-pole bandpass filter with a Butterworth response is:

$$\rho = n \left(20 \log \frac{Q_u + Q_L}{Q_u}\right) \tag{1}$$

where

ρ = filter insertion loss (db)

Q_u = Unloaded cavity Q

Q_L = Loaded cavity Q

n = number of filter poles

now

$$Q_u = \frac{Q_L Q_u'}{Q_L + Q_u'} \tag{2}$$

where

Q_L = tuning capacitor Q

Q_L = 1,000 (Refer to Fig. 11)

and $$Q'_u = 120 \times s \times \sqrt{f_o} \qquad (3)$$

where

S = cavity diameter (inches)

f_o = center frequency of resonance (MH)

$$Q'_u = 120 \times 1 \times \sqrt{449}$$

$$Q'_u = 2540$$

then

$$Q_u = \frac{2540\ (1000)}{1000 + 2540} = 716$$

$$Q_L = \frac{f_o}{BW_{3db}} \qquad (4)$$

where

BW_{3db} = 3db bandwidth of each cavity section under loaded conditions.

$$Q_L \simeq \frac{449}{28} = 16$$

thus

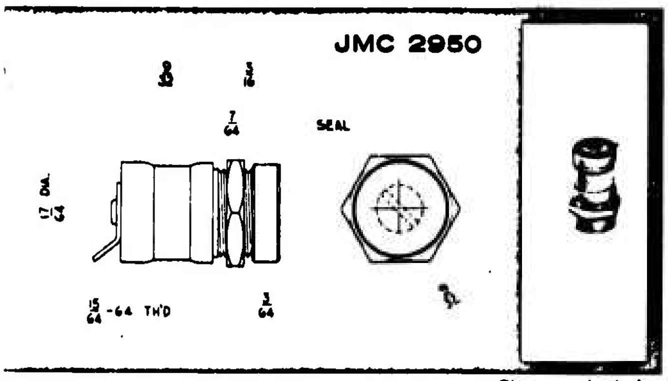
$$\rho = 8\ (20 \log \frac{716 + 16}{716})$$

$$\rho = 8\ (0.172)$$

$$\rho = 1.37 \text{ db}$$

2950

"MINIATURE" SERIES



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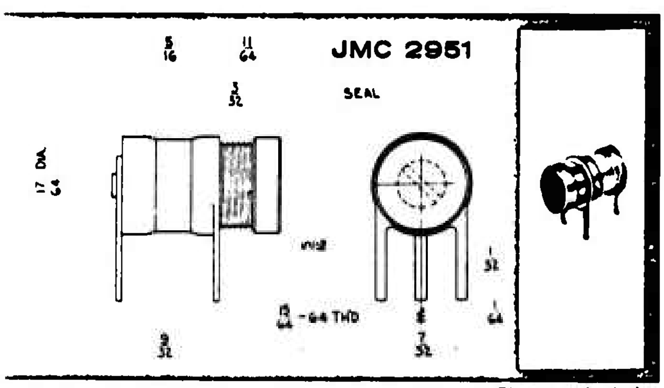

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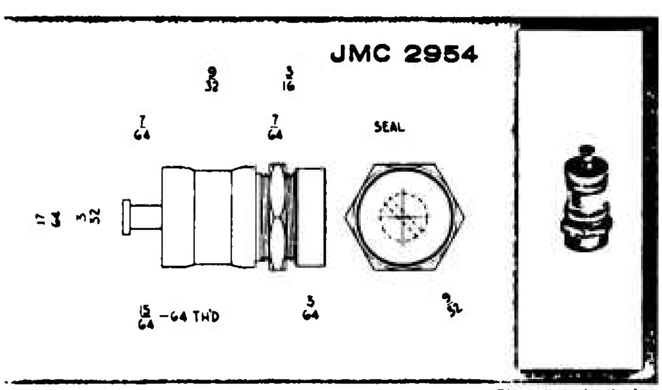

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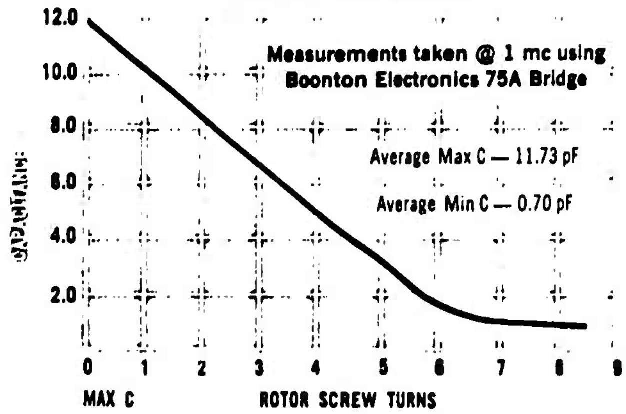
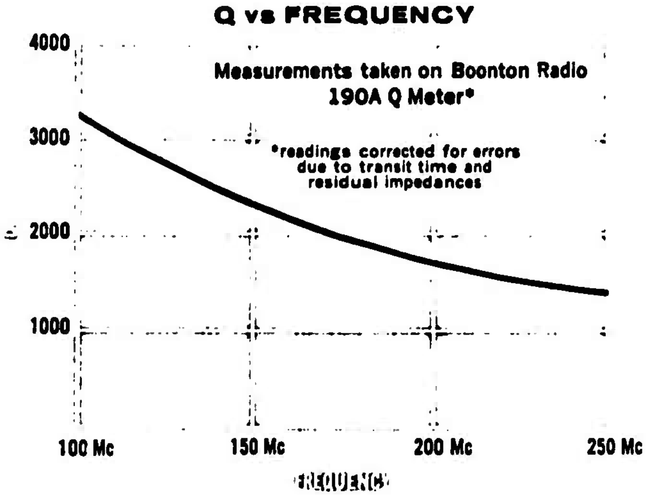
Figure 11

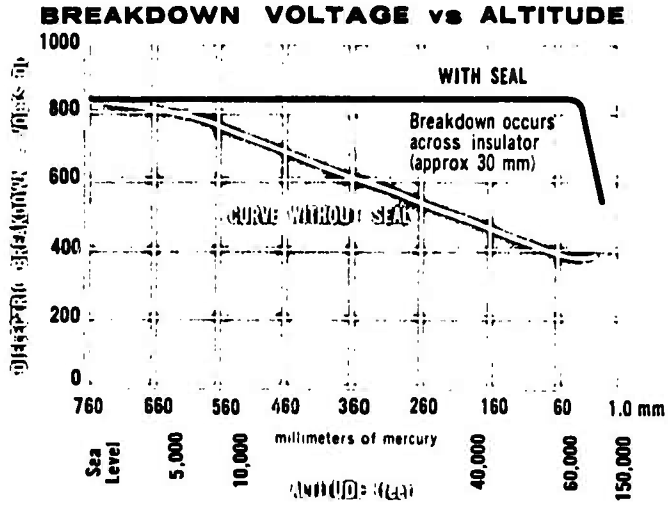
Δ C vs ROTATION



Q vs FREQUENCY



BREAKDOWN VOLTAGE vs ALTITUDE



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Measured values of the post filter insertion loss are 1.0 db. The additional loss due to diplexing with the receiver preselector was negligible. This is considerably better than calculated. Special tuning techniques were developed at ITTFL to obtain this performance.

The transmitter output power selection is accomplished by two simple steps, (1) a resistor value change (located external to the power supply module) to adjust the +28 volt power supply voltage, (2) slight retuning of the transmitter module. Since the power supply efficiency was designed to be fairly independent of the load, the transponder primary input power reduces in almost direct proportion to a corresponding reduction in output power.

2.1.3 Phase Shift (PD para. 3.1.3)

The SECOR System is a phase measuring system and as such, the transponder must impart the smallest possible phase shift to the incoming ranging subcarriers over wide ranges of signal input and conditions of environment as referenced in paragraph 3.4.3 of the purchase description.

In distance measuring systems, where range information is derived by comparing the phase of a modulating signal (or signals) of the transmitted carrier with that of the detected signal(s) of the received carrier, phase stability is of primary importance. The method used to meet the phase stability requirements for the MATS transponder is called Phase Following Feedback.

Although we can find the literature covering the subject of PFFB, sometimes called "Frequency compressive feedback," or "frequency following demodulation," it may be helpful to analytically review the technique and in so doing, point out basic design areas of concern.

A simplified block diagram of the PFFB loop used in the MATS transponder is shown in Figure 12. Important operational parameters are shown. An assumption to be made for the following expressions is that the bandpass elements used in the transponder have a flat amplitude response and linear

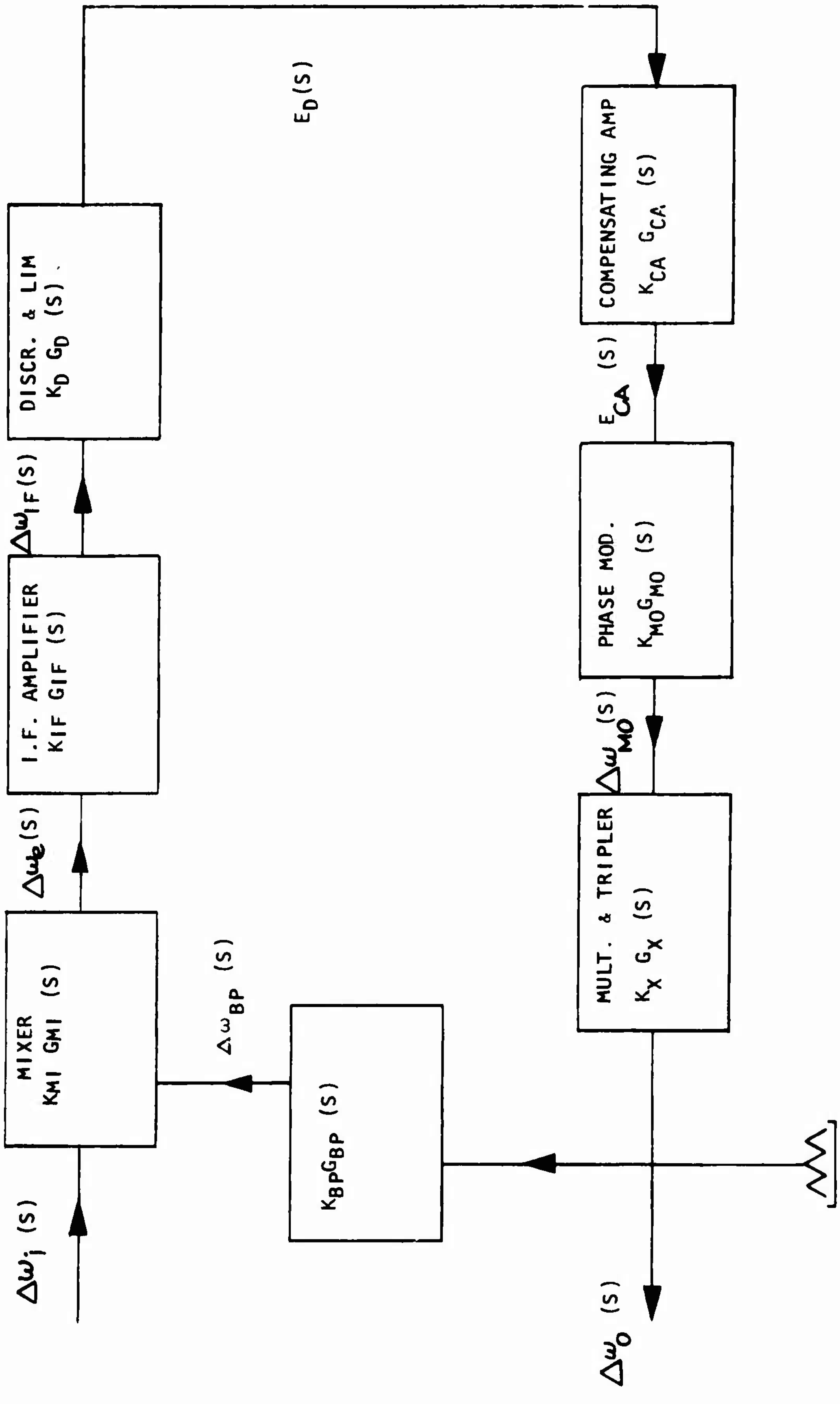


Figure 12. Block Diagram of Transponder for Transfer Function Analysis

phase shift in the frequency region which includes all signficant sidebands so that a linear analysis does apply. This requirement is not overly stringent since these characteristics are essential for good inherent phase stability and low modulation distortion.

The following equations are derived from the block diagram:

$$\frac{\Delta \omega_o(S)}{\Delta \omega_i(S)} = \frac{K_{FO}\, G_{FO}\,(S)}{K_T\, G_T\,(S) + 1} \tag{1}$$

$$K_{FO}\, G_{FO} \quad K_{MI}\, K_{IF}\, K_D\, K_{CA}\, K_{MO}\, K_X\, G_{MI}\, G_{IF}\, G_D\, G_{CA}\, G_{MO}\, G_X$$

$\dfrac{\Delta \omega_o S}{\Delta \omega_i (S)}$ closed loop transfer function

$\dfrac{\Delta \omega_o (S)}{\Delta \omega_e (S)}$ Transfer function of forward loop (2)

$$K_T\, G_T\,(S) \quad K_{FO}\, G_{FO}\,(S) \cdot K_{BP}\, G_{BP}\,(S)$$

$\dfrac{\Delta \omega_{BP} (S)}{\Delta \omega_e (S)}$ Transfer function of total loop (3)

where:

$\Delta \omega_i(S)$ Laplace Transform of the frequency deviation of the input carrier frequency

$\Delta \omega_{BP}(S)$ Laplace Transform of the frequency deviation of the output carrier frequency, after passing through B.P. filter

$\Delta\omega_e(S)$ = Laplace Transform of the frequency deviation of the input intermediate frequency

K_{MI} = Frequency independent factor of the transfer function (in this case, $K_{MI} = 1$)

$G_{MI}(S)$ = Frequency dependent factor of the transfer function of the modulation frequencies in the mixer.

K_{IF} = Frequency independent factor of the transfer function = 1

$G_{IF}(S)$ = Frequency dependent factor of the transfer function of the modulation frequencies in the I.F. amplifier

K_D = Gain of the phase detector

$G_D(S)$ = Frequency dependent factor acting upon the modulation frequencies in the phase detector

$G_{CA}(S)$ = Frequency dependent factors acting upon the modulation frequencies in the compensating amplifier.

K_{MO} = Gain of the phase modulator in radians phase shift per volt of modulation signal

$G_{MO}(S)$ = Frequency dependent factor acting upon the modulation frequencies in the phase modulator

K_x = Frequency multiplication factor

$G_X(S)$ = Frequency dependent factor acting upon the modulation frequencies in the multipliers and transmitter.

K_{BP} = Frequency independent factor of the transfer function = 1

$G_{BP}(S)$ = Frequency dependent factor acting upon the modulation frequencies in the Band Pass filter.

K_{CA} = Gain of the compensating amplifier.

Let total loop gain at any frequency $j\omega$ be

$$K_T G_T (j\omega) = K_{\mu\beta} e^{j(\alpha_\mu + \alpha_\beta)} \tag{4}$$

and

$$K_{FO} G_{FO} (j\omega) = K_\mu e^{j\alpha_\mu} \tag{5}$$

where

$K_{\mu\beta}$ = system loop gain = $P K_D K_{CA} K_{MO} K_X$

P = input modulation angular frequency

K = forward loop system gain

Therefore equation (1) reduces to

$$\frac{\Delta\omega_o (j\omega)}{\Delta\omega_i (j\omega)} = \frac{K_\mu e^{j\alpha_\mu}}{K_{\mu\beta} e^{j(\alpha_\mu + \alpha_\beta)} + 1} \tag{6}$$

We can equate the closed loop transfer function $\frac{\Delta\omega_o (j\omega)}{\Delta\omega_i (j\omega)}$ to a modulation index change since

$$m = \frac{\Delta\omega}{\omega_m}$$

then

$$\frac{\Delta\omega_o (j\omega)}{\Delta\omega_i (j\omega)} = \frac{m_o}{m_i} e^{j\phi_o} \tag{7}$$

where

m_o = transponder output modulation index

m_i = transponder input modulation index

Letting $\tan \alpha_o = \dfrac{K_{\mu\beta} \sin(\alpha_\mu + \alpha_\beta)}{K_{\mu\beta} \cos(\alpha_\mu + \alpha_\beta) - 1}$

and $R = \sqrt{K_{\mu\beta}^2 - 2K_{\mu\beta} \cos(\alpha_\mu + \alpha_\beta) + 1}$

then $$\frac{m_o}{m_i} = \frac{K_\mu e^{j\alpha}}{R e^{j\alpha_\gamma}}$$

$$\frac{m_o}{m_i} = \frac{K_\mu}{\sqrt{1 - 2K_{\mu\beta} \cos(\alpha_\mu + \alpha_\beta) + K_{\mu\beta}^2}} \tag{8}$$

and

$$\alpha_o = \alpha_\mu - \tan^{-1} \frac{K_{\mu\beta} \sin(\alpha_\mu + \alpha_\beta)}{K_{\mu\beta} \cos(\alpha_\mu + \alpha_\beta) - 1} \tag{9}$$

since $K_{BP} = 1$ then $K_\mu = K_{\mu\beta}$ (our case). Of particular interest is to note that in equation (8) as $K_{\mu\beta}$, the system loop gain, becomes large (i.e. $K_{\mu\beta} >> 1$) then $m_o \rightarrow mi$, that is, the retransmitted modulation index approaches that received by the transponder.

For example taking MATS transponder parameters, $K_{\mu\beta} = 30$, and $(\alpha_\mu + \alpha_\beta)$ a small angle, $K_\mu = K_{\mu\beta}$, then

$$\frac{m_o}{m_i} \cong \frac{K_{\mu\beta}}{K_{\mu\beta} + 1} = \frac{30}{31}$$

or

$$m_o \cong m_i$$

The output modulation index m_0 is very closely equal to m_i, the input modulation index. Also noting equation (9) and allowing the practical case (our's) of

$$\alpha_\beta = \text{very small}$$

$$\alpha_\mu >> \alpha_\beta$$

$$K_{\mu\beta} >> 1$$

then $$\alpha_o = \tan^{-1} \frac{K_{\mu\beta} \sin(\alpha_\mu + \alpha_\beta)}{K_{\mu\beta} \cos(\alpha_\mu + \alpha_\beta) - 1}$$

can be written

$$\alpha_o = \alpha_\mu - \tan^{-1} \frac{K_{\mu\beta} \sin \alpha_\mu}{K_{\mu\beta} \cos \alpha_\mu - 1}$$

$$= \alpha_\mu - \tan^{-1} \left(\frac{\sin \alpha_\mu}{\cos \alpha_\mu} + \frac{K_{\mu\beta} \sin \alpha_\mu}{-1} \right)$$

$$= \alpha_\mu - \tan^{-1} \left(\tan \alpha_\mu - K_{\mu\beta} \sin \alpha_\mu \right)$$

$$= \alpha_\mu - \alpha_\mu + \tan^{-1} K_{\mu\beta} \sin \alpha_\mu$$

$$\alpha_o = \tan^{-1} K_{\mu\beta} \sin \alpha_\mu \tag{11}$$

As $K_{\mu\beta} \uparrow$. Then $\alpha_o \longrightarrow 0$

Since we are interested in $\Delta\alpha_o$ with conditions such as temperature and dynamic range which in themselves create loop phase shifts, then we calculate $\frac{d\alpha_o}{d\alpha_i}$, $\frac{d\alpha_o}{d\alpha_\beta}$. These quantities representing the rate of change of the output modulation signal phase shift with changes in the forward loop and feedback loop phase shifts. In addition, we are also interested in $\frac{d\alpha_o}{dK_{\mu\beta}}$ which represents the rate of change of the output phase shift with open loop gain.

From equation (7)

$$\frac{d\alpha_o}{d\alpha_\mu} = \frac{1 - K_{\mu\beta} \cos(\alpha_\mu + \alpha_\beta)}{K_{\mu\beta}^2 - 2K_{\mu\beta} \cos(\alpha_\mu + \alpha_\beta) + 1} \qquad (12)$$

$$\frac{d\alpha_o}{d\alpha_\beta} = \frac{K_{\mu\beta} \cos(\alpha_\mu + \alpha_\beta) - K_{\mu\beta}^2}{K_{\mu\beta}^2 - 2K_{\mu\beta} \cos(\alpha_\mu + \alpha_\beta) + 1} \qquad (13)$$

Figures 13 and 14 give plots of these quantities with $K_{\mu\beta}$ as a parameter. Notice that in the limit for $(\alpha_\mu + \alpha_\beta) = n\pi$ where n = 0, 2, 4, etc. and $K_{\mu\beta} >> 1$ we have

$$\frac{d\alpha_o}{d\alpha_\mu} \longrightarrow -\frac{1}{K_{\mu\beta}} \quad \text{and} \quad \frac{d\alpha_o}{d\alpha_\beta} = -1$$

thus, a change in the forward path phase is reduced at the transponder output by the factor $\frac{1}{K_{\mu\beta}}$, but a change of phase in the feedback path is not reduced by PFFB. For the MATS transponder, we have

$$K_{\mu\beta} = 30$$

$$\Delta\alpha_\mu = 60^\circ \qquad \text{(Worse case temperature change)}$$

$$\Delta\alpha_\beta < 0.1^\circ$$

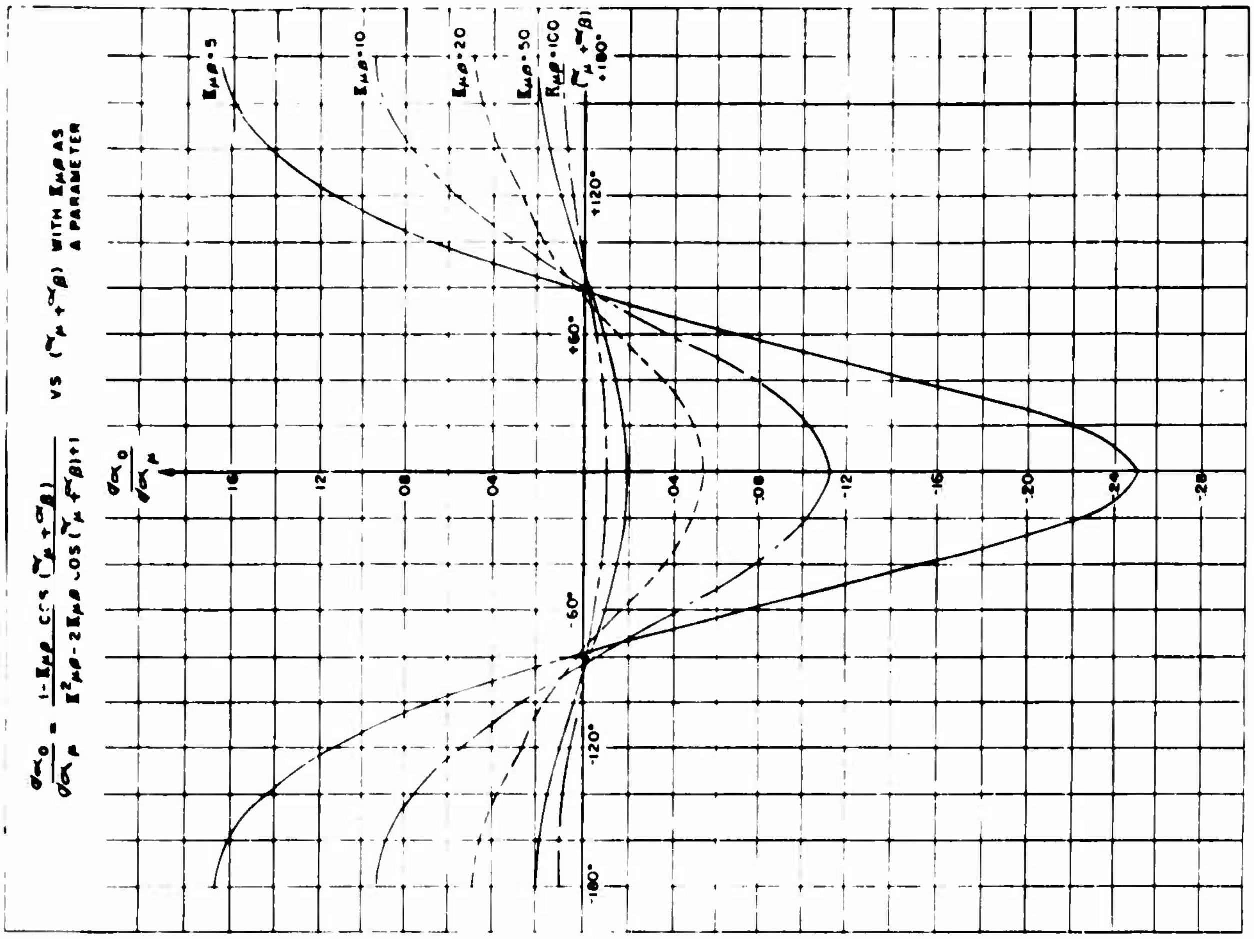


Figure 13

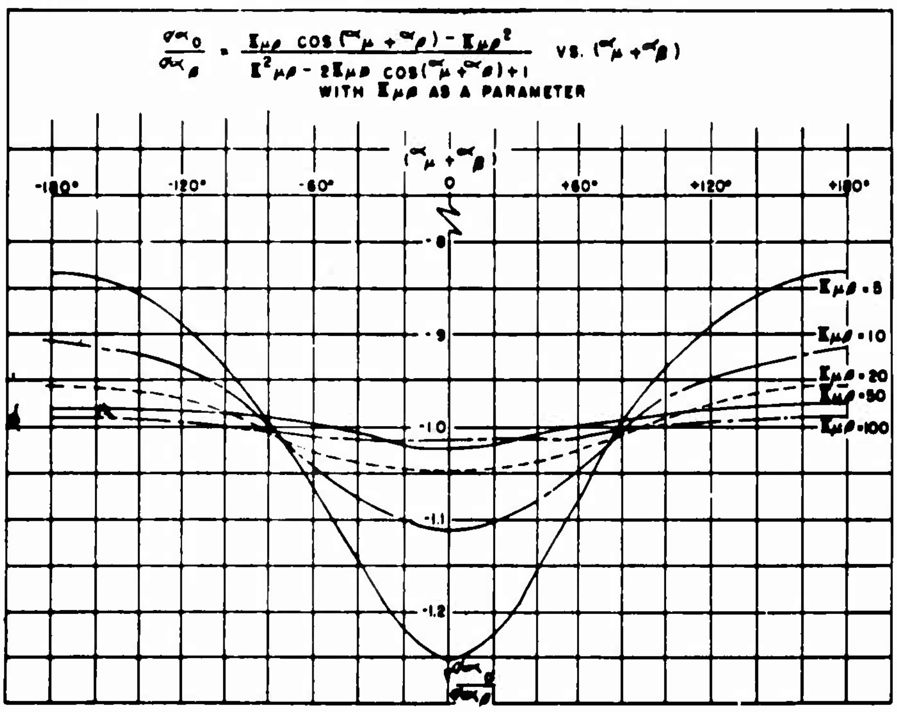


Figure 14

$$\frac{d\alpha_o}{d\alpha_\mu} \cong \frac{1}{K_{\mu\beta}}$$

$$\Delta\alpha_o = \frac{\Delta\alpha_\mu}{K_{\mu\beta}} = \frac{60^\circ}{30} = 2^\circ$$

$$\frac{d\alpha_o}{d\alpha_\beta} \cong -1$$

$$\Delta\alpha_o = -\Delta\alpha_\beta = -0.1^\circ$$

It becomes evident from the above, that a good PFFB transponder should exhibit a negligible change in phase in the feedback path since PFFB at its best, will not correct this phase change. Indeed, the MATS transponder feedback path contains a negligible number of wideband components in this path, all of which exhibit $< 0.1^\circ$ phase change over all environmental and operating conditions.

One should also note that, assuming the above condition is satisfied $\Delta\alpha_\beta << \Delta\alpha_\mu$, the closed loop phase change is proportional to the forward loop phase change. Thus, $\Delta\alpha_\mu$ should be kept as small as possible.

From equation (7) we have

$$\frac{d\alpha_o}{dK_{\mu\beta}} = \frac{\sin(\alpha_\mu + \alpha_\beta)}{K_{\mu\beta}^2 - 2K_{\mu\beta}\cos(\alpha_\mu + \alpha_\beta) + 1} \tag{14}$$

Figure 15 shows a plot of $\frac{d\alpha_o}{dK_{\mu\beta}}$ vs. parameter $(\alpha_\mu + \alpha_\beta)$ the total phase shift of the loop.

These plots are used in the choice of the open loop gain $K_{\mu\beta}$, and the phase $\alpha_\mu + \alpha_\beta$. Since it is desirable to minimize the phase variation with variations in circuit parameters, the following conclusions are reached: (1) The loop gain $K_{\mu\beta}$ should be maintained as high as feasible. System parameters dictate the maximum practical gain that can be achieved with reasonable hardware.

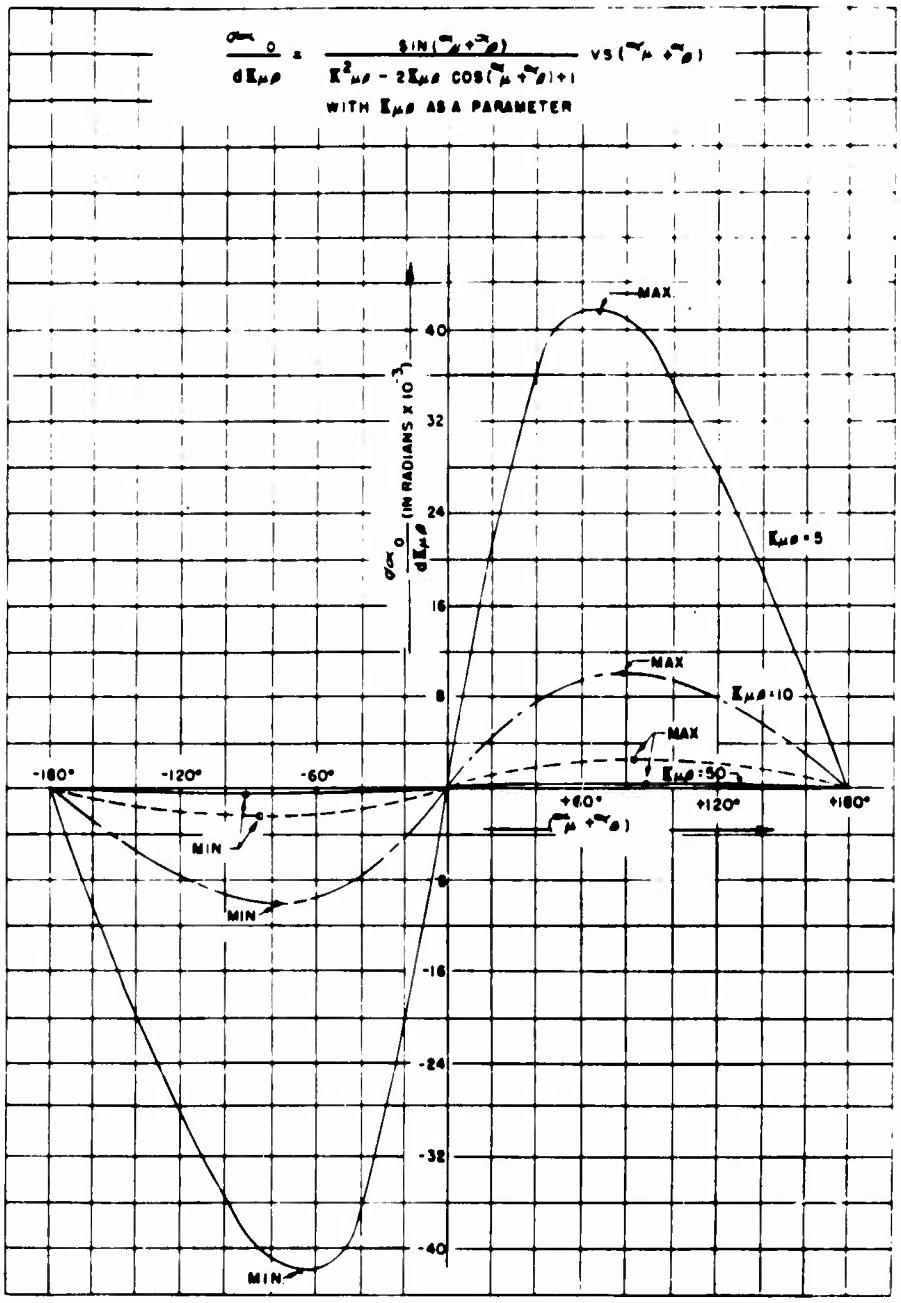

WITH $K_{\mu\phi}$ AS A PARAMETER
MAX
MAX
MAX
MIN
MIN
MIN
$K_{\mu\phi}=5$
$K_{\mu\phi}=10$
$K_{\mu\phi}=50$
40
32
24
16
8
-8
-16
-24
-32
-40
-180°
-120°
-60°
+60°
+120°
+180°

Figure 15

(2) Since $\frac{d\alpha_o}{d\alpha_\mu}$ and $\frac{d\alpha_o}{dK_{\mu\beta}}$ vs. $(\alpha_\mu + \alpha_\beta)$ are approximately 90^o out of phase, examination of the actual expected variations of α_μ and $K_{\mu\beta}$ is needed to choose the value of $(\alpha_\mu + \alpha_\beta)$ at which the sum of both effects is minimized.

Another factor to be considered, however, is to provide a closed loop system with adequate stability margins. This dictates the maximum allowable loop gain $K_{\mu\beta}$ and also the selection of $\alpha_\mu + \alpha_\beta$ near to $2\pi n$ where n = 0, 1, 2, 3 . . . In general, a value near zero for $\alpha_\mu + \alpha_\beta$ best satisfies both requirements.[4]

In the MATS transponder, the stability of the loop dictates $(\alpha_\mu + \alpha_\beta) = 2\pi n$ where n = 0, 1, 2, 3, - - - .

since $$\alpha_\mu >> \alpha_\beta$$

then $$\alpha_\mu = 2\pi n$$

and $$\frac{d\alpha_o}{dK_{\mu\beta}} \cong 0$$

that is, the closed loop output phase is minimally affected by changes in the open loop system gain.

Also, one should note that although $K_{\mu\beta}$ would normally be chosen as large as possible to minimize $\Delta\alpha_o$, another primary factor, receiver sensitivity, dictates a maximum $K_{\mu\beta}$ since practical filter bandwidths at the MATS subcarrier frequencies are limited.

MATS crystal filters necessitate the use of type DD cuts which exhibit high level spurious at 60% of their center frequencies. Consideration of all of the above factors yielded $K_{\mu\beta}$ maximum of 30.

[4]Note: The mixer is a phase subtractor and, therefore, represents π radians of the total $(\alpha_\mu + \alpha_\beta)$. $\therefore$ $(\alpha_\mu + \alpha_\beta) = \pi n$, where n = 1,3,5, . . . for MATS.

2.1.4 Solid State Components (PD para. 3.1.4)

The transponder is designed using silicon solid state components exclusively. This choice assured reliable operation over the required ambient temperature range (-4°F to +160°F) and a sufficient margin to allow extended temperatures to be used if deemed justified at some future date.

2.1.5 Materials (PD para. 3.2)

All materials used in the transponder are of such substance as not to deteriorate in the environment (vacuum, radiation, vibration and heat) to which the transponder will normally be subjected and defined in the purchase specifications. Any questionable materials were tested within the expected environment prior to inclusion within the final design.

2.1.6 General (PD para. 3.3.1)

The transponder is designed for a useful life of at least one year's normal operation. The transponder is to be used in satellite configurations in orbits up to 2500 nautical miles at inclinations from 0 to 90 degrees.

The transponder is designed to operate in a "standby", "receive" or "transmit" condition. In the "standby" condition (minimum power mode), only those circuits necessary to place the transponder in a "receive" condition upon receipt of a coherent carrier are operative. In the "receive" mode those circuits required to provide access to the select call and other normal command signals, are engaged. Upon receipt of a select call subcarrier the "transmit" mode is initiated. In the "transmit" condition, all circuits are energized and the transponder is capable of performing in a manner called out in modified purchase description.

The one year operation is satisfied by using worst case circuit design techniques. This assures us that if all components of a particular network

were to attain the worst possible parameter variation as defined by their individual specs, then the circuit would still perform its function satisfactorily.

In addition, the networks in themselves were conservatively designed, when possible, to allow beyond spec limits to occur (except in a few isolated cases) and still operate satisfactorily. High reliability components were used whenever possible.

During "standby" operation, a minimal number of circuits are operative consistent with the necessity of commanding the transponder into a receive mode. A coherent carrier commands the transponder into the receive mode which can be used for commands to telemetry or the select call operation. A select call command initiates full power to all transponder circuits.

The select call frequency is easily changed to any subcarrier frequency within the 400 to 600 KHz range (and beyond if necessary).

2.1.7 Weight (PD para. 3.3.2)

The weight of the transponder, including interconnecting cables, connectors, and hardware, is 11-1/2 pounds.

The transponder weight is kept to a minimum by (1) the lack of a main frame assembly, (2) the scalloping the modules to remove that metal not necessary for structural strength and (3) the minimum use of covers between modules resulting from the use of the backs of each as the RF shield for those adjacent.

Magnesium was considered, but it was too expensive and difficult to work.

2.1.8 Size and Shape (PD para. 3.3.3)

The overall volume of the transponder, including mounting brackets, connectors and hardware, is 235 cubic inches for all power levels. The transponder is housed in one (1) regular figured rectangular package. The outside

dimensions, excluding mounting brackets, mating connectors and other hardware are 4 1/4" x 6 1/2" x 8 1/2". (See Figure 3)

A pictorial of the complete package is shown in Figure 16. All available space within the allowable maximum dimensions was used to the fullest extent possible due to the high volume of circuitry found necessary to satisfy all the transponder functional requirements.

This new package was developed to satisfy the MATS configuration, and in addition, (1) to allow easy component accessibility by both engineering and technical personnel during both operating and nonoperating conditions and (2) easy module interchangeability.

Note that the transponder can be unfolded like a book as shown in Figure 16, allowing almost complete access to all circuitry without removal of a single wire. Thus the transponder can be in operating condition when completely exposed. Alignment of the overall transponder is thus simplified and tailor values can easily replace variables so that no adjustments are provided, nor needed, once the alignment is complete (except those necessary for retuning the transmitter to different output power levels). Inter - module wiring (except coax) is hard wired point-to-point to eliminate connectors.

2.1.9 Dissimilar Metals (PD para. 3.3.4)

Dissimilar metals are not used in intimate contact unless protected against electrolytic corrosion. Dissimilar metal combinations and comparable metal coupling is as defined in MIL-STD-33586.

The modules are aluminum and circuitry boards copper. Both are gold-plated. All screws are stainless steel.

2.1.10 Connectors (PD para. 3.3.5)

The use of connectors is minimized to the highest degree practical. Where connectors are required, they are keyed or positioned so that mating

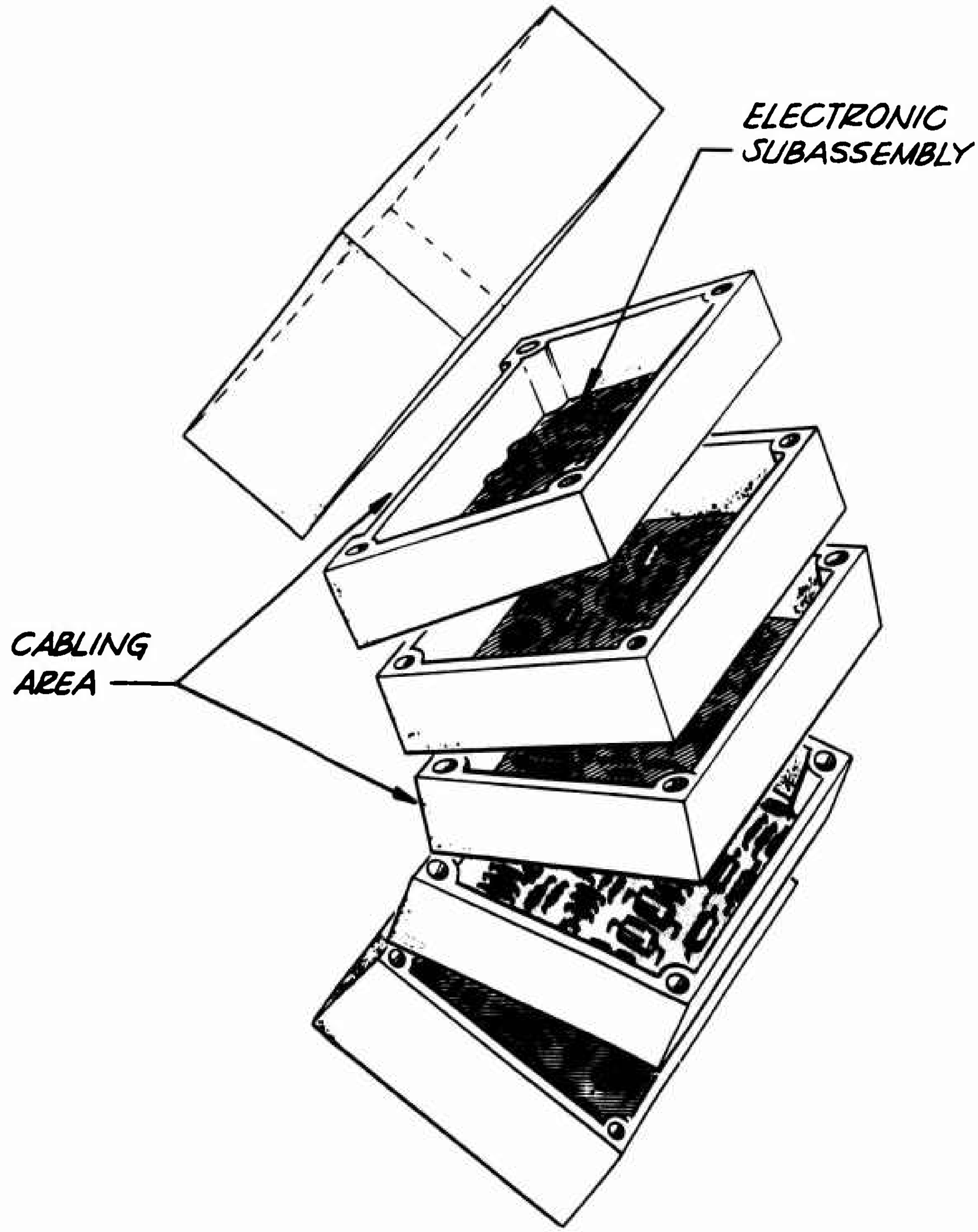


Figure 16

Transponder Modules

errors cannot be made. All connectors, once mated, are capable of being locked in place either by screw or other positive technique.

The transponder input/output connectors are (1) 2-telemetry multipin-keyed and screw lock types, (2) 1-power supply primary source connector-keyed and screw lock type, (3) 2-RF connectors single conductor positive lock types. All RF interconnections between modules use single conductor positive lock types.

2.1.11 Environment (PD para. 3.3.6)

2.1.11.1 Thermal Vacuum (PD para. 3.3.6.1)

The transponder performs within the limits called out in paragraph 3.4 of purchase description in thermal environments from -4°F to +160°F in vacuums at least 1×10^{-5} mm of mercury. In addition, the transponder is capable of being stored in a vacuum of at least 1×10^{-5} mm of mercury at a temperature of -30°F without damage to the transponder. Once removed, the transponder is capable of performing within the limits called out in paragraph 3.4.

The use of components capable of withstanding the thermal vacuum requirements is essential. All active components (i.e., transistors, integrated circuits), are capable of at least -55°C to +125°C operating temperatures and minimum storage temperatures equal to or exceeding this range. For example, the 2N918 transistor and the μA702A integrated circuit are used extensively throughout the transponder. Refer to Table 1.

TABLE 1

	2N918	μ A702A
Max Storage Temperature	-65°C to +300°C	-65°C to +150°C
Maximum Operating Temperature	-55°C to +200°C	-55°C to +125°C

All passive components, resistors, capacitors, diodes, etc. are chosen, such that the normal operating thermal environment is conservative compared to the component ratings. All diodes are silicon, capacitors are either mica DM's, ceramic CK's or tantalytic SCM's, and resistors 1/4 watt-carbon film conservatively derated to less than 25% of maximum ratings.

2.1.11.2 Vibration (PD para. 3.3.6.2)

The transponder performs within the limits called out in paragraph 3.4 purchase description after being subjected to the following types and levels of vibration.

a. Sinusoidal Vibration (all major perpendicular axes)

Frequency (cps)	Applied Vibration Level (zero to peak acceleration)
5 - 14	0.5 in. DA
14 - 40	±5.0 g
40 - 50	±7.5 g
50 - 70	±30 g
70 - 2000	±22.0 g
2000 - 3000	±20.0 g

b. Random Vibration (all major perpendicular axes)

Frequency (cps)	Applied Vibration Level (g^2/cps)
5 - 14	0.07 g^2/cps
15 - 50	0.10 g^2/cps
50 - 200	0.40 g^2/cps
200 - 2000	0.20 g^2/cps

2.1.11.3 Shock (PD para. 3.3.6.3)

The transponder performs within the limits called out in paragraph 3.4 of purchase description after being subjected to three half-sine wave shock impulses of 0.5 milliseconds duration at a level of 200 g's in each major perpendicular axis.

2.1.11.4 Acceleration (PD para. 3.3.6.4)

The transponder performs within the limits called out in paragraph 3.4 of purchase description after being subjected to sustained acceleration forces of at least 22 g's for periods of at least 15 minutes in all major perpendicular axes.

To meet the vibration, shock, and acceleration requirements of this design, the following areas of the transponder have received special consideration:

- All internal cabling and wiring have been spot-bonded wherever possible to prevent excessive movement.
- All circuit board components have been conformally coated to the boards to eliminate movement and dampen vibration. Also, components that are exceptionally susceptible to damage have been bonded to the cases or circuit boards when practical.
- All cavity filters or networks using helical coils have been steel wire wrapped to their terminations so they can easily withstand the vibration requirements.
- All circuit boards have been designed with the components, terminals and board representing a solid, non-flexing piece of hardware.
- The individual system chassis have been designed with rigid outside frames and solid center webbs to provide maximum strength at the component mounting locations.
- All of the individual system chassis are held together at each corner with adequate hardware to form a solid assembly.

Precision machined module surfaces insure minimum movement between individual module mounting surfaces.

2.1.12 Primary Power Requirements (PD para. 3.3.7)

The transponder performs within the limits called out in paragraph 3.4 of purchase description when subjected to input voltage variations from 11.5 to 17.5 VDC. The input power requirements of the transponder shall not exceed those listed below, for the mode shown, with 17.5 volts DC applied at its primary power input.

Standby Mode	0.8 watts maximum (average)
1.5 watt mode	23.0 watts maximum
3.5 watt mode	35.0 watts maximum
4.5 watt mode	39.0 watts maximum

The standby input power is 1.2 watts if the transponder is "on" 100% of the satellite orbit. A patent pending technique used by ITTFL to provide duty cycle switching of the standby supply can reduce the average power dissipation in standby to 0.25 watts. The prototype transponders are adjusted to a 50% duty cycle for an average power dissipation of 0.8 watts.

The input powers for various output powers are as shown above. Although the 4.5 watt mode input power is within the modified purchase description requirements, the 3.5 watt and 1.5 watt modes were high. The primary reasons for this are:

(1) 28 volt supply voltage from transponder power supply not adjustable over 25 to 33 volts rather than design limits of 20 to 30 volts.

(2) The fixed receiver loss is independent of the transmitter power, but the spec requirement is proportional to it.

As previously noted in paragraph 2.1.2 of this report, the diplexer designed for this program is special and indeed advances the state of the art in the area of insertion loss for a given number of filter sections. The design of this diplexer was a primary consideration in meeting the state of the art power efficiency of the transponder.

Other primary factors affecting the amount of primary power used (during transmit) by the transponder are: (1) the required output power; (2) the transmitter efficiency, primarily the high power stages, and; (3) the power supply (DC-to-DC converter) efficiency.

The required output power for the 4.5 watt mode is 4.5 watts at 449 MHz, and 4.5 watts at 224.5 MHz. Since the diplexer insertion loss is 1.0 db and the 224.5 MHz post filter insertion loss is 0.5 db, then the required output power from the 445 MHz final amplifier is 1.26 x 4.5 watts = 5.65 watts, and correspondingly, 5.05 watts from the 224.5 MHz final amplifier. The 449 MHz amplifier is best designed about class B or C operation to ensure optimum efficiency.

The collector circuit efficiency, η, is defined as the ratio of the a-c power delivered to the real load, R_L, divided by the d-c input power. When an ideal unilateral device is operated class B or class C, η will be greater than 78 percent. However, in a practical situation, collector-circuit efficiency will be either artificially high or much lower.

The principal causes of difficulty in obtaining ideal class B operation in a common-emitter circuit are excess emitter lead inductance, L_e, and internal capacitance, C_{cb}, between the collector and base terminals that shunt the active region of the transistor. The extra capacitance and its associated time constant with the resistances it drives produces a base drive that tends to keep the transistor ON when the external drive is moving it toward cutoff. L_e also tends to keep emitter current flowing, rather than allowing the emitter to be sharply cut off. The result of these effects is to produce a larger conduction angle and lower efficiency. With careful design, reasonable approximations to class B waveforms can be achieved with efficiencies of 50 to 70 percent.

The transducer gain provides an effective measure of circuit performance. Transducer gain is defined as the actual power output to the real load divided by the power available from the source. When this quantity is large compared to 1, the collector circuit

efficiency will be a meaningful quantity. When transducer gain drops significantly, appreciable power flow from input to output through passive elements may be occurring and the collector circuit efficiency then will be an artificial measure of circuit performance. This is obvious when it is realized that a transducer gain of 1 can be achieved with a passive matching network requiring no d-c input power. The normal definition of η would make it infinite for this case. For low-gain transistors, efficiency would better be defined as transistor efficiency.

$$\eta_{transistor} = \frac{P_{AC\ OUT}}{P_{AC\ IN} + P_{DC\ IN}} \times 100\%$$

Figure 17 contains a curve showing the RF power output vs. RF power input at 400 MHz for the 3TE 440 transistor. This device was found to provide the most gain of any device presently available for the required power output at 449 MHz. Other transistors investigated were 2N4040 and RCA 2N3375. For the typical collector of circuit efficiency of 65 percent published in the 3TE440 spec (slightly optimistic for our case at 449 MHz), we have

$$P_{DC\ IN\ FINAL} = \left[\frac{P_{AC\ OUT}}{\eta_{TRANS.}} - P_{AC\ IN} \right]$$

$$P_{AC\ OUT} = 5.65 \text{ WATTS}$$

$$\eta_{TRANS.} = 0.65 \text{ or } 65\%$$

$$P_{AC\ IN} = 1.9 \text{ WATTS}$$

$$\therefore \quad P_{DC\ IN\ FINAL} = \left[\frac{5.65}{0.65} - 1.9 \right] = 6.8 \text{ WATTS}$$

Repeating the above for the 1.9 watt driver 3TE 450. Refer to Figure 18 for 3TE450 curve.

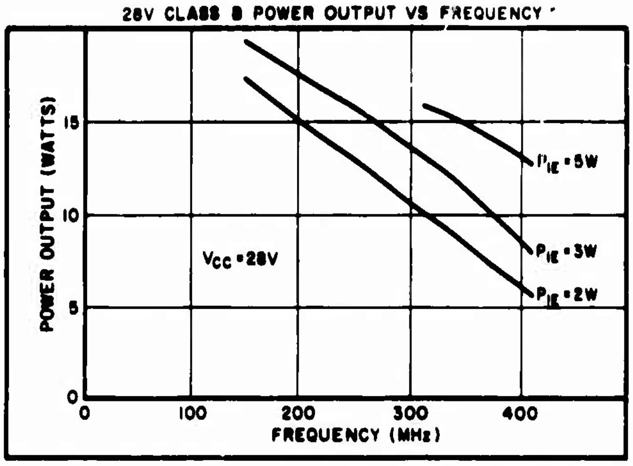
$$P_{\substack{DC\ IN \\ 1ST\ DRIVER \\ DOUBLER}} = \left[\frac{1.9}{\eta_{(50\%)\ DOUBLER}} - 0.5 \right] = 3.3 \text{ WATTS}$$

$$P_{\substack{DC\ IN \\ 1ST\ DRIVER \\ AT\ 224.5\ MH}} = \left[\frac{0.5}{0.65} \right] \cong 0.77 \text{ WATTS}$$

$$P_{\substack{DC \\ 224.5\ FINAL}} = \left[\frac{5.05}{0.65} - 1.0 \right] = 6.7 \text{ WATTS}$$

$$P_{\substack{DC \\ 224.5\ DRIVER \\ DOUBLER}} \cong \left[\frac{1}{0.5} \right] = 2.0 \text{ WATTS}$$

3TE 440



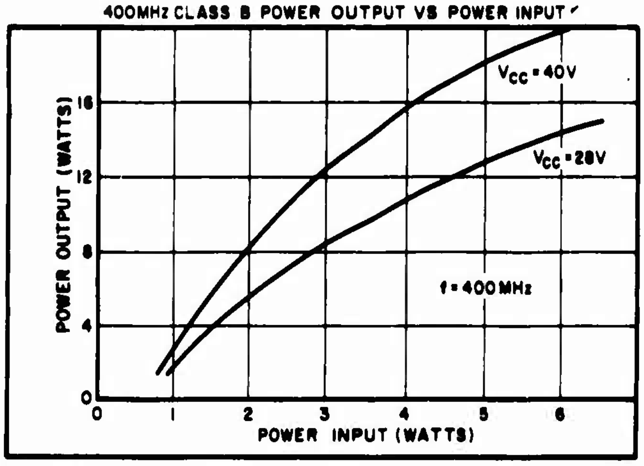
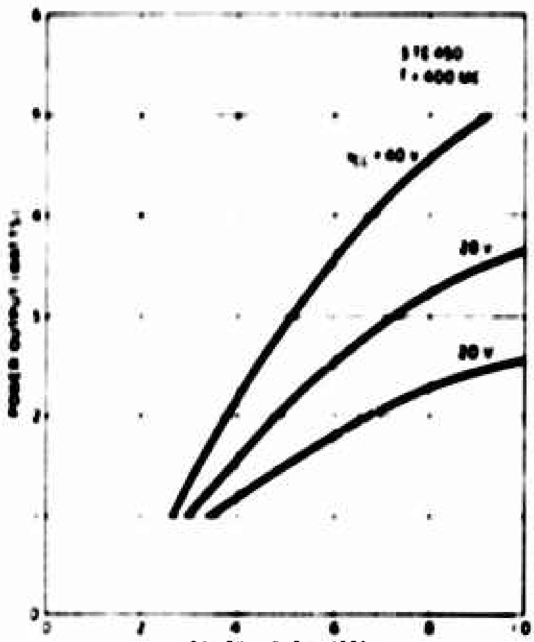


Figure 17



TYPICAL 400 Mc POWER OUTPUT
vs POWER INPUT

Figure 18

Now

$$P_{\substack{DC\ IN \\ TOTAL}} = P_{\substack{DC\ IN \\ 449\ FINAL}} + P_{\substack{DC\ IN \\ 449\ 1ST \\ DRIVER \\ DOUBLER}} + P_{\substack{DC\ IN \\ 224.5\ 2ND \\ DRIVER \\ DOUBLER}} + P_{\substack{DC\ IN \\ 224.5 \\ FINAL}}$$

$$+ P_{\substack{DC\ IN \\ 224.5\ DRIVER}} + \left[P^{*}_{DC\ IN} \right]$$

$P^{*}_{DC\ IN}$ = input power to operate all other transponder circuits = 4 WATTS

$$\therefore \quad P_{\substack{DC\ IN \\ TOTAL}} = 6.8 + 3.3 + 0.77 + 6.7 + 3.0 + 4$$

$$P_{\substack{DC\ IN \\ TOTAL}} = 24.57 \text{ WATTS}$$

Since the maximum primary power allowed in the 4.5 watt mode is 39 watts, then the transponder power supply efficiency would have to be:

$$\eta_{\substack{POWER \\ SUPPLY}} = \frac{24.57}{39} \times 100\% = 63\%$$

Notice that in the above calculations, near optimum conditions were assumed: (1) nearly theoretical efficiencies; (2) no allowances for temperature effects; (3) the use of the same driver for 449 MHz and 224.5 MHz finals, and; (4) perfect VSWR matching (i.e., VSWR = 1). An overall DC-to-DC converter efficiency of 70% minimum is required in order to meet the power input requirement.

Since the power supply efficiency is reasonably independent of the load, the input power level required at lower RF power output levels decreases almost in direct proportion.

The transponder was designed to operate with a minimum of primary power in the "standby" mode. The specification requires the maximum standby primary power not to exceed 1 watt. Table 2 indicates the standby power distribution of the transponder on a module basis.

The RF amplifiers, KMC 2N3880 transistors, were chosen on the primary basis of lowest noise figure available at 421 MHz. As common base RF amplifiers they exhibit very stable electrical characteristics, and low noise figures, but require ≈ 70 mw of input power.

The modulator/oscillator module supplies a standby L.O. to the receiver first mixer. The required stable L.O. is derived by multiplication (X24) from an 18.7 MHz reference carrier oscillator. The 150 milliwatts dissipation is the result of an extensive power reduction program. The X6 step recovery diode provides the theoretical 1/n efficiency criteria (where n is the multiplication ratio). Since the hybrid isolator has a 3.5 db insertion loss, and the first mixer requires $\approx$ 0 dbm injection, then 75 MWs for the 75 MHz amplifier, 50 MWs for the X4 multiplier which includes losses in a 75 MHz 2-pole filter to reject the 18 MHz fundamental, and 25 MWs for a TCXO results in a reasonable power dissipation for the module.

The data amplifier/demodulator module provides a phase lock loop to track and detect the transmitted carrier, a correlation loop to command the power supply to the operation mode and the 11.225 MHz reference oscillator. This TCXO is again 25 MWs, and all 2N918 amplifiers are starved to 2.0 ma, (i.e., the lowest acceptable value consistent with worst case design practice). The DC amplifiers are integrated circuits μA 702As. Unless ICs are specifically designed for MATS standby voltages (which would be quite prohibitive based upon the small quantities used) then this 40-50 MW circuit type is the best that can be presently expected. The IF module provides $\approx$120 dbm of power gain under standby conditions to ensure that all effects of the environment will not

TABLE 2

MATS POWER REQUIREMENTS

S.T.B.Y.	DIP (ma)	I-F (ma)	Demod. (ma)	Mod/Osc. (ma)	Total (ma)	Secondary Power (watts)	Primary Power (watts)
+6V	7.0	38	12.5	13.5	71.0	.426	
-6V	6.5	18	20	11.0	55.5	.333	
Total	13.5	56	32.5	24.5	126.5	.759	1.26

degrade the performance of the transponder below the minimum required to bring it out of a standby condition. Thus, a power dissipation of $\approx$ 25 MWs/10 db gain is reasonable since the wide bandwidth required by each IF stage ($\approx$ 15 MHz) for a constant gain bandwidth product, allow only a small effective gain per stage. In addition to the above, frequency conversion, limiting, the VCO, etc. are also powered in the IF module.

The best power supply efficiency attainable under standby condition is 60 to 65%. The efficiency of the power supply is limited by the large primary voltage variation and the high degree of regulation of the output voltages required by the transponder. The resultant 0.8 watt input power is thus the result of a duty cycle mode of standby operation.

2.1.13 Radio Frequency Interference (PD para. 3.3.8)

The transponder meets the EMI requirements as defined in MIL-STD-826. This is basically a requirement to provide filtering at the transponder output, the end result of which provides > 60 db attenuation to all frequencies which are harmonically related to those transmitted and $\geq$ 80 db attenuation if non-harmonically related. The 8-pole, 449 MHz post filter provides $\approx$ 48 db/oct attenuation to all frequencies outside of the 3 db bandwidth and thus easily rejects all unwanted frequencies. The 2-pole 224.5 MHz post filter, providing a 12 db/oct slope, will reject the second harmonic 449 MHz by a minimum of 64 db. Sub-harmonic frequencies (related to 18.7 MHz fundamental crystal), are rejected by filters within the multiplication chain.

Radiation from other areas of the transponder should be minimal since precautions such as (1) providing RF shielded cables for all interconnections external to the modules, (2) completely enclosing the cables, (3) completely enclosed machined modules (providing excellent electrostatic shielding), (4) complete control of module ground locations to minimize the effect of electro-magnetic radiation, and (5) shields grounded at both connector ends.

2.1.14 "Select Call" Circuit (PD para. 3.3.9)

Circuitry is provided within the transponder to recognize whether or not a "select call" signal is present and carry out the operation required to place the transponder in a "transmit" condition. The circuit is capable of being readily changed and fixed (by minor tuning and crystal insertion, etc.) to recognize a given subcarrier within the range of 400 to 600 KHz. The bandpass of the circuitry associated with the "select call" frequency is designed so that the -45 dbm to -115 dbm or other spurious transponder signals can not falsely trigger the transponder in a "normal" operate condition. The above holds true for all worst combinations of environment and voltage variations. Provisions are made to override the "select call" circuit from an external switching function without having to insert a "select call" signal at the receiver input. The connections for the override feature terminate at an easily accessible point on an external connector.

The original specification requires a "select call" operation with composite subcarrier signals. This operation is not compatible with a high modulation index phase following type of transponder since proper control of the composite subcarriers by the ground stations is required to initiate the "select call" operation.

When the PFFB loop is not operative, that is, when the transponder transmitter is not operating, high index composite subcarriers entering the receiver can exceed the dynamic range of the phase detector. In addition, when passing through an index of, say, 2.4 radians, the receiver phase lock loop can lose carrier lock since theoretically no carrier power exists at this particular index.

A simple replacement of the "select call" crystal filter is required to change the subcarrier frequency anywhere in the 400 to 600 KHz range or beyond.

The "select call" crystal filter is multiple pole, thus providing a skirt selectivity of ≥ 12 db/octave.

Subcarrier frequencies out of band are attenuated reducing the probability of falsely triggering the transponder into the transmit mode.

2.1.15 Transmitter Output (PD para. 3.3.10)

The transponder transmitter is capable of providing either 1.5, 3.5 or 4.5 watts output from the transmitter terminals, which includes the diplexer, on both 224.500 MHz and 449.000 MHz. Choice of the desired output necessitates the selection of a power supply resistor, and minor tuning of the transmitter. Minor tuning and calibration is defined as the process required by a technician not thoroughly familiar with the transponder, but by reading instructions, to tune and calibrate the transponder in a two-hour period using standard laboratory equipment for normal operation and performance as defined in other paragraphs of this document.

The transmitter system power output change is accomplished by: (1) changing the B+ voltage to the transmitter from 28 VDC to lower settings as defined in the alignment procedure and accomplished by changing a single resistor, located external to the power supply module, and; (2) slight retuning of the transmitter module. Variable capacitors are provided with adjustments made using standard alignment tools.

2.1.16 Antenna Input/Output (PD para. 3.3.11)

The transponder requires, at a maximum, two (2) antenna connections for reception of signals from the ground complex and transmission of the ranging and timing data back to the ground complex. A diplexer is provided as an integral part of the transponder, prior to the antenna output terminations, to allow reception of signals on 420.9375 MHz and the transmission of signals on 449.000 MHz. All output terminals, including the diplexer, are capable of handling either the 1.5, 3.5 or 4.5 watt output configuration.

The input/output impedance of all antenna connections is 50 ohms over the total bandwidth of interest. The VSWR does not exceed 1.5 for the 449 or 224.5 MHz ports.

The sensitive receiver employed in the transponder has demanded an unusual degree of filtering at the receiver input and the transmitter output. The filters selected are both **eight-pole.** Tuning of the diplexer, including both filters, is performed as an integrated unit prior to integration with the transmitter and receiver. The resultant losses are as shown in Figure 19.

It is impractical to check the VSWR of the transmitter outputs over the bandwidth since this would require driving the high power stages at frequencies widely separated from 449 MHz or 224 MHz, thus providing, in some cases, reactive loads, at full power, sufficient to exceed device dissipation. Therefore, the band of interest, as far as VSWR measurements, is taken to be <u>center frequency only</u>.

The VSWR, as applied to the transponder receiver is 2.0:1, since power match denoted by a VSWR of 1 does not provide a minimum receiver noise factor and thus, maximum transponder sensitivity.

2.1.17 <u>Bandpass</u> (PD para. 3.3.12)

The overall bandpass of the transponder and data channels (including all subcarrier commands), is designed to minimize phase shift, as defined in paragraph 3.4.3 of the purchase description. The design considerations include all system instabilities, any combination of modulation index, S/N, the effects of doppler and all other requirements defined in this document. Some of the system considerations are:

(1) the maximum radial velocity of the satellite which contains the transponder, is that associated with an elliptical orbit whose perigee is 200 N. miles and whose apogee is 2000 N. miles in altitude;

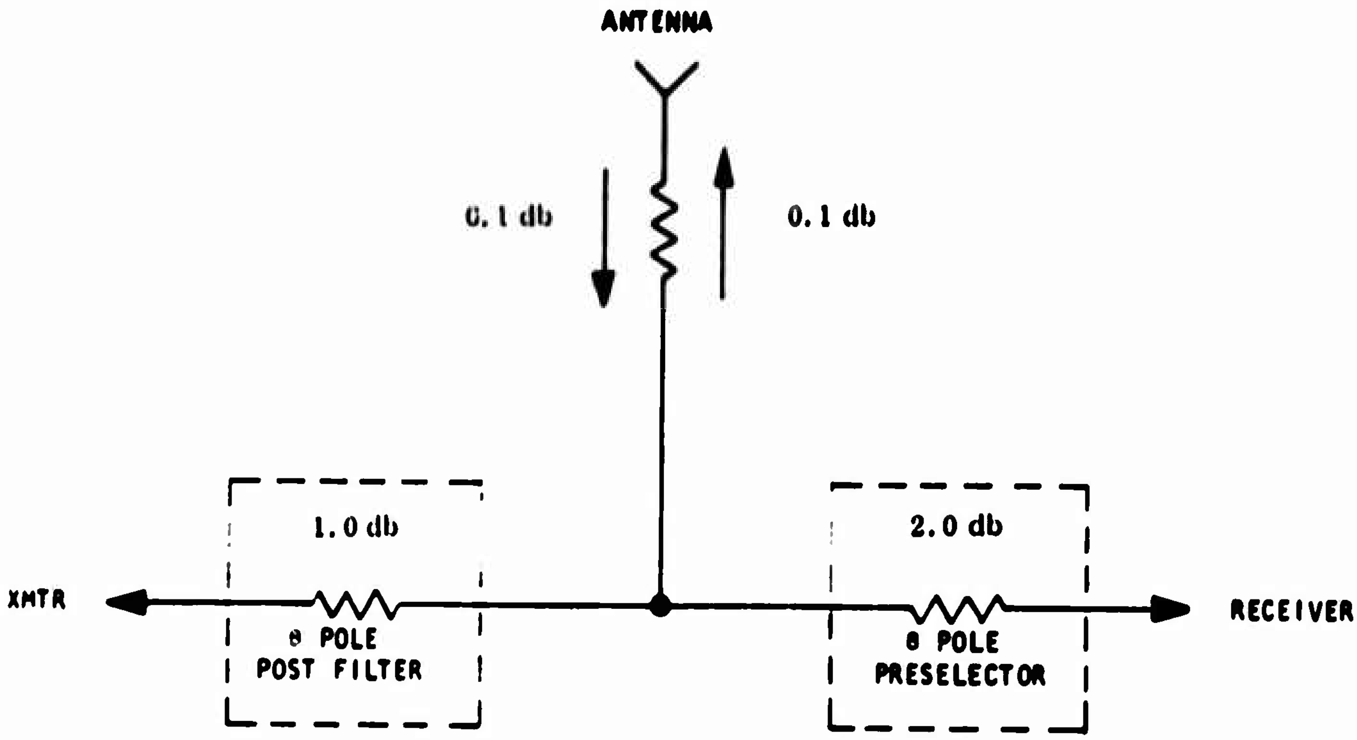

ANTENNA
0.1 db
0.1 db
1.0 db
XMTR
8 POLE
POST FILTER
2.0 db
RECEIVER
8 POLE
PRESELECTOR

Figure 19

(2) Frequency	Stability
a. 420.9375 MHz carrier	0.001%
b. All subcarriers	0.005%

(3) modulation indexes of the individual ranging or timing subcarriers lie within the range of 0.5 to 2.5 radian; (4) the transponder ranging data signal-to-noise ratio is +12 db as defined in paragraph 3.4.11 of the purchase description.

Analysis derives the closed loop performance of the transponder as related to open loop parameters. The primary open loop parameters of importance being $\Delta\alpha\mu$, $\Delta\alpha\beta$ and $K\mu\beta$.

Refer to the MATS transfer function Block Diagram, Figure 12 and compare it to the functional Block Diagram, Figure 1. Since the takeoff point for PFFB occurs at the output of the 449 MHz transmitter final amplifier, the only circuit incorporated into the feedback path is a 2-pole cavity filter. This filter's primary function is the elimination of undesirable L. O. frequencies, mainly the 224.5 MHz. Calculating envelope time delay associated with this filter, we have:

$$t \text{ envelope} = \frac{\Delta O(\omega) \text{ radians}}{\Delta\omega \text{ radians}} (n) \qquad (1)$$

where t envelope = envelope delay of the bandpass network.

n = number of poles in the network

$\Delta O(\omega)$ = 90^{o} linear phase shift between 3 db points for each pole.

= 1.57 radians/90^{o}

$\Delta\omega$ = 3 db bandwidth in radians for each pole.

therefore $$t_{\text{envelope feedback path}} = \frac{1.57}{50 \times 10^{6} (2\pi)} \times 2 = .01 \ \mu\text{sec}$$

Although cavity filters exhibit time delay stabilities much better than 10%, for argument's sake the benefit of the doubt is given the reader. Thus

$$\Delta t_{\substack{\text{feedback}\\ \text{path}}} = .1 \times .01 = .001\ \mu\text{sec}$$

which at 600 KHz (i.e., the worst case highest modulating frequency), yields

$$T_{600\ \text{KHz}} = 1.67\ \mu\text{sec}/360^{\circ}$$

and

$$\Delta\alpha_{\beta} = \frac{(.001)\ (360)}{1.67} = 0.215\ \text{degrees}$$

for

$$K_{\mu\beta} = 30$$

then

$$\frac{\Delta\alpha O}{\Delta\alpha_{\beta}} \cong -1$$

$$\Delta\alpha O \Big|_{\Delta\alpha\beta} \cong -0.215^{\circ}$$

Thus, the output phase change exhibited by the transponder due to $\Delta\alpha\beta$ is minimal.

Table 3 summarizes the results of a similar analysis for the forward loop portion of the transponder on a module basis. The changes in envelope delay were arbitrarily taken as 10% for each module (except the crystals). This is typical of the performance exhibited by the breadboard transponder over complete environment and signal level changes related to the S/N ratio and modulation index variables. Without the crystal filters we have

$$\Delta' \chi_u = 7.52$$

noting that

$$\Delta \chi_{u\,\text{crystals}} = 60^{\circ}$$

and

$$\Delta \chi_u = \Delta \chi_u\,(\text{crystals}) + \Delta \chi'_u$$

$$\Delta \chi_u = 67.5^{\circ}$$

since

$$\Delta \chi_{u\,\text{crystals}} \gg \Delta \chi'_u$$

then $$\Delta \chi_u \cong \Delta \chi_{u\,\text{crystals}}$$

Thus the choice of crystals becomes an all important parameter in the design of a phase stable transponder. Since a DD type crystal cut is the most temperature stable as compared to the other alternatives of DT, and CA, this type was chosen. An unfortunate byproduct of its superior temperature stability is spurious products at undesirable frequencies of sufficient magnitudes to cause loop stability problems. Considerable effort was devoted in the demodulator

TABLE 3

Module	(usec) Envelope Delay	(usec) Change in Delay	(degrees) Change in Phase
IF	0.136	0.0136	1.75
Data Amp/Demod (Excluding Crystals)	0.0312	< 0.00312	0.67
Phase Mod/Osc	0.207	<0.02	4.3
Transmitter	0.037	<0.0037	0.8
Crystal Filters	2500.0	----	60.0

design to negate the spurious effects. The results allowing the use of a crystal with $\Delta\alpha_{\mu\,\text{crystal}} = 60^{\circ}$ max. Since most of the phase shift occurs between 23°F and -4°F, it is reasonable to expect the $\Delta\alpha_o$ performance outside of this temperature range to be correspondingly improved.

Since
$$\frac{\Delta\alpha_o}{\Delta\alpha_\mu} = \frac{1}{K_\mu Q} = \frac{1}{30}$$

then
$$\Delta\alpha_o'|_{\Delta\alpha_\mu} = 67.5/30 = 2.25^{\circ}$$

Some circuitry exists outisde of both the forward and feedback paths, namely the diplexer, RF amplifiers and associated filters. Refer to Table 4 for calculation results. The total $\Delta\alpha_o$ with respect to these circuits is

$$\Delta\alpha_{o\,\text{out of loop circuits}} = 2.5^{\circ}$$

Notice that the Diplexer filters were allowed a 10% change in delay. Again we are being ultra conservative since, for cavity type filters, the actual delay change is <1%.

TABLE 4

CIRCUIT	ENVELOPE DELAY (usec)	CHANGE IN DELAY (usec)	CHANGE IN PHASE (Degrees)
Pre Selector	.05	.005	1.07
Post Filter	.05	.005	1.07
RF Amplifier	.0063	.00063	.136
Other Filters	.0083	.00083	.178

Thus, from the above we can calculate the $\Delta\alpha_o$.

$$\Delta\alpha_{o\,\substack{\text{calculated}\\ \text{worse case}}} = \Delta\alpha_{o\,\substack{\text{out of}\\ \text{loop circuits}}} + \Delta\alpha_{o\,\substack{\text{respect to}\\ \text{forward loop}}} + \Delta\alpha_{o\,\substack{\text{respect to}\\ \text{feedback loop}}} \qquad (2)$$

$$= 2.5^{\circ} + \frac{67.5}{30} + 0.215$$

$$= 2.5 + 2.25 + .215$$

$$= 5^{\circ}$$

In practice, with the breadboard transponder, it has been found that $\Delta\phi_o / \Delta\phi_{cr}$ has been by far the predominant factor and thus equation (2) can be rewritten

$$\Delta\phi_{o\,\mathrm{actual}} \approx \Delta\phi_o \Big|_{\Delta\phi_{cr}}$$

or even more simply, and almost as accurately

$$\Delta\phi_{o\,\mathrm{actual}} \approx \Delta\phi_o \Big|_{\Delta\phi_{cr\,\mathrm{crystal}}} \qquad (3)$$

denoting that the choice of this crystal filter is of major importance to the design of a phase stable transponder, assuming all other factors, due to careful design, can be neglected. This case applies to the MATS transponder and is all important to its future potential.

As far as those parameters relating to frequency stabilities and associated orbital parameters, they are used primarily for specifying the phase lock loop parameters along with the stabilities of those oscillators associated with the transponder itself.

A most unique problem is that of a carrier acquisition. Because the satellite is moving past four ground stations at geographically different locations, and with slightly differing RF carrier frequencies, the transponder must be prepared to perform the carrier acquisition process anew on each received pulse in the SECOR transmission frame sequence. Thus, in a period of 50 ms, the transponder will encounter four different carrier frequencies with four different doppler offsets. Acquisition and de-acquisition must be virtually instantaneous.

The system designed to meet the requirements of the purchase specification is shown in the block diagram, Figure 1. The system consists basically of two loops in parallel. One is the carrier phase locked loop, and the other is the phase following feedback loop. They are not, however, independent.

The gain of the carrier loop is a function of carrier level, and as such, is a function of the feedback gain of the PFFB loop. Further, the detection of the subcarriers is done in the phase detector by product detection. The output of the product detector is given by:

$$E_0 = k \sin(\theta_{dc} + \theta_1 \cos p_1^{\,t} + \theta_2 \cos p_2^{\,t} + \ldots) . \qquad (4)$$

For $\theta_{dc} + \theta_1 + \theta_2 << 1$ radian, the output would be linear and the output levels would be independent. It is evident that in analyzing loop stability (i.e. $\theta \rightarrow 90$ degrees) the two loops must be treated as one.

Rather than treat the problem of interdependence analytically, we chose to present a subjective analysis. This is justified since the nonlinear analysis required to solve the problem is not adequately solved at this time.

Let us take two practical cases: Case I, high S/N ratio in both loops, Case II low S/N condition within the PFFB loop.

Case I: High S/N ratio ensures that the signal power is >> noise power within the PFFB loop. For $K_{\mu\varphi} = 30$ we have

$$m_{pd} = \frac{m_i}{1 + K_{\mu\varphi}} \qquad (5)$$

where

m_{pd} = modulation index presented to phase detector

m_i = modulation index received by transponder

solving for highest m_{pd} we have

$$m_{pd} = \frac{15 \text{ radians Peak}}{1 + 30} = 0.484 \ \underline{\text{radians peak}}$$

Thus, a linear condition prevails within the phase detector and the PFFB and carrier loops are for all practical purposes independent.

Case II: Low S/N ratio is defined as that level in which the noise power within the PFFB loop is sufficient to momentarily prevent the loop from tracking the phase of the incoming modulation. As the tracking error increases, equation (35) becomes

$$m_{pd} = \frac{m_i \epsilon}{1 + K_{\mu\varrho}}$$

where

ϵ = factor relating to the PFFB tracking error.

$\epsilon \rightarrow 1$ for negligible error

$\epsilon \rightarrow (1 + K_{\mu\varrho})$ as the tracking error approaches $\geq 90^o$.

As the loop S/N ratio approaches threshold and below

$$m_{pd} \rightarrow m_i$$

and the phase detector output can no longer linearly detect the modulation. Since carrier level is a function of the modulation index and the PLL gain is a function of carrier input level, then one can easily accept the fact that the carrier loop can lose lock for weak transponder input signals.

The basic requirements for the system to be sequentially interrogated by four ground stations in 10 ms bursts with 2.5 ms spacing between bursts, establishes the following operating criteria for the transponder:

(1) The phase locked loop must initially acquire to turn the transmitter on.

(2) The phase locked loop must reacquire every 12.5 ms upon a different carrier frequency.

(3) All transient responses of the previous station must die down in 2.5 ms.

Since several hundred milliseconds are allotted for initial transmitter turn-on, and the time delay for turn-off is at least seven seconds, the first requirement does not impose any difficulty. The main concern in this area is to set the select call threshold sufficiently high to ensure a low probability of false alarm turn-on due to noise peaks. Diodes, in conjunction with filter bandwidths wide enough to pass only the highest select call modulation rate (i.e., 20 pps with a 15% duty cycle), satisfy the requirement.

The need for reacquisition every 12.5 ms, and the transient response requirement impose some difficulty, unless the transmitted signal is properly shaped to allow the feedback loops to respond. A recommended signal shape would be that shown in Figure 20. The carrier alone would be on for 0.5 ms; then the modulation would be applied gradually for the next 2.5 ms. Full modulation would exist for an additional 5.5 ms, and then the modulation level would drop off to zero in 2.5 ms with the carrier off after another 0.5 ms.

The bandwidth of the phase locked loop must be sufficient to acquire the carrier in the 0.5 ms under all conditions of possible carrier frequency uncertainty. This uncertainty is due to doppler, oscillator instabilities, and dc offsets in the voltage controlled oscillator loop. They are as follows:

• Doppler uncertainty	±12 KHz
• Received Signal Frequency 0.001% Stability at 421 MHz	± 4.2 KHz
• VCO Stability	± 3.9 KHz
• Reference oscillator stability	± 1.6 KHz
• Local Oscillator Instability 0.0005% at 449 MHz	± 2.25 KHz
• Phase Detector DC Offset ±1.5 mv at 15 KHz/volt	± 4.5 KHz

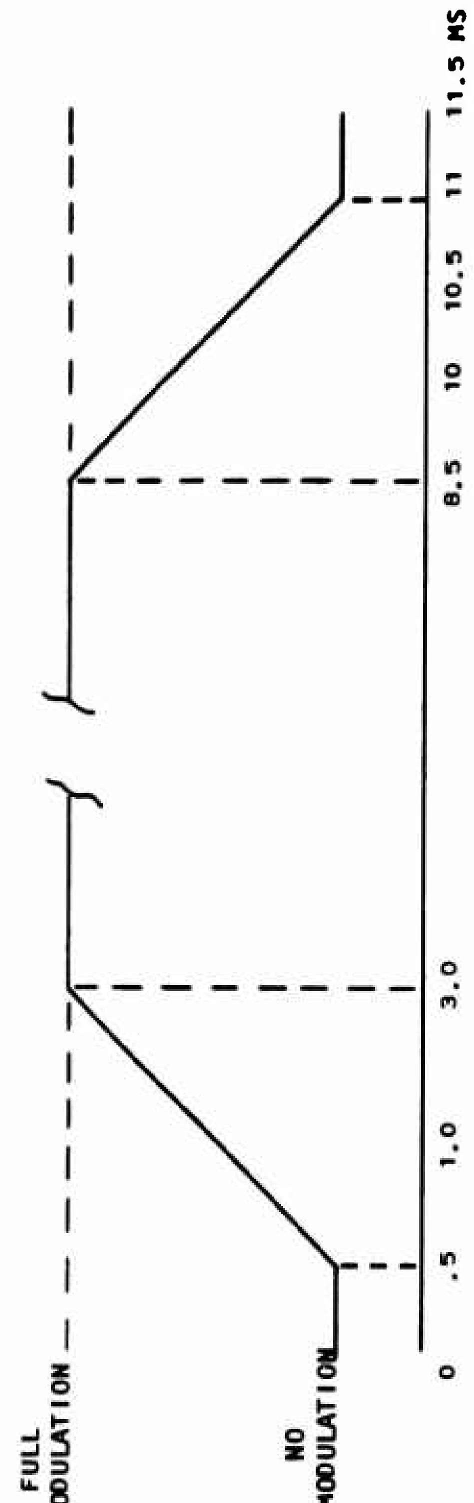


Figure 20. Interrogator Transmitted Modulation Waveform

The maximum error would be approximately +28.5 KHz. The time required for carrier lock is

$$t_L \geq \frac{K\,(\Delta f)^2}{f_{nn}^3}$$

where f_{nn} is the lowpass bandwidth of the loop, Δf is the frequency uncertainty, and the constant K is a function of the damping constant of the loop, equal to approximately 4 for a minimum noise bandwidth loop.

f_{nn} = 30 KC (Refer to Figure 14 for open loop Bode plot)

K = 4 (for a 0.7 damping factor; MATS loop)

Δf = 28.5 KHz

$$t_L \geq \frac{4\,(28.5 \times 10^3)^2}{(30 \times 10^3)^3} \geq 116 \text{ usec}$$

Thus, the allowable lock-up time of 0.5 usecs assures carrier acquisition. One should note that the above lock time is modified somewhat by the constantly changing loop bandwidth f_{nn}.

The bandwidth change occurs since the loop gain can be as much as 15 db less after acquisition than before, (Refer to Figure 21), due to coherent AGC action.

f_{nn} after acquisition = 5 KHz

The S/N of the loop for an input signal of -115 dbm before acquisition is

S/N = 115 - KT + B + F

where

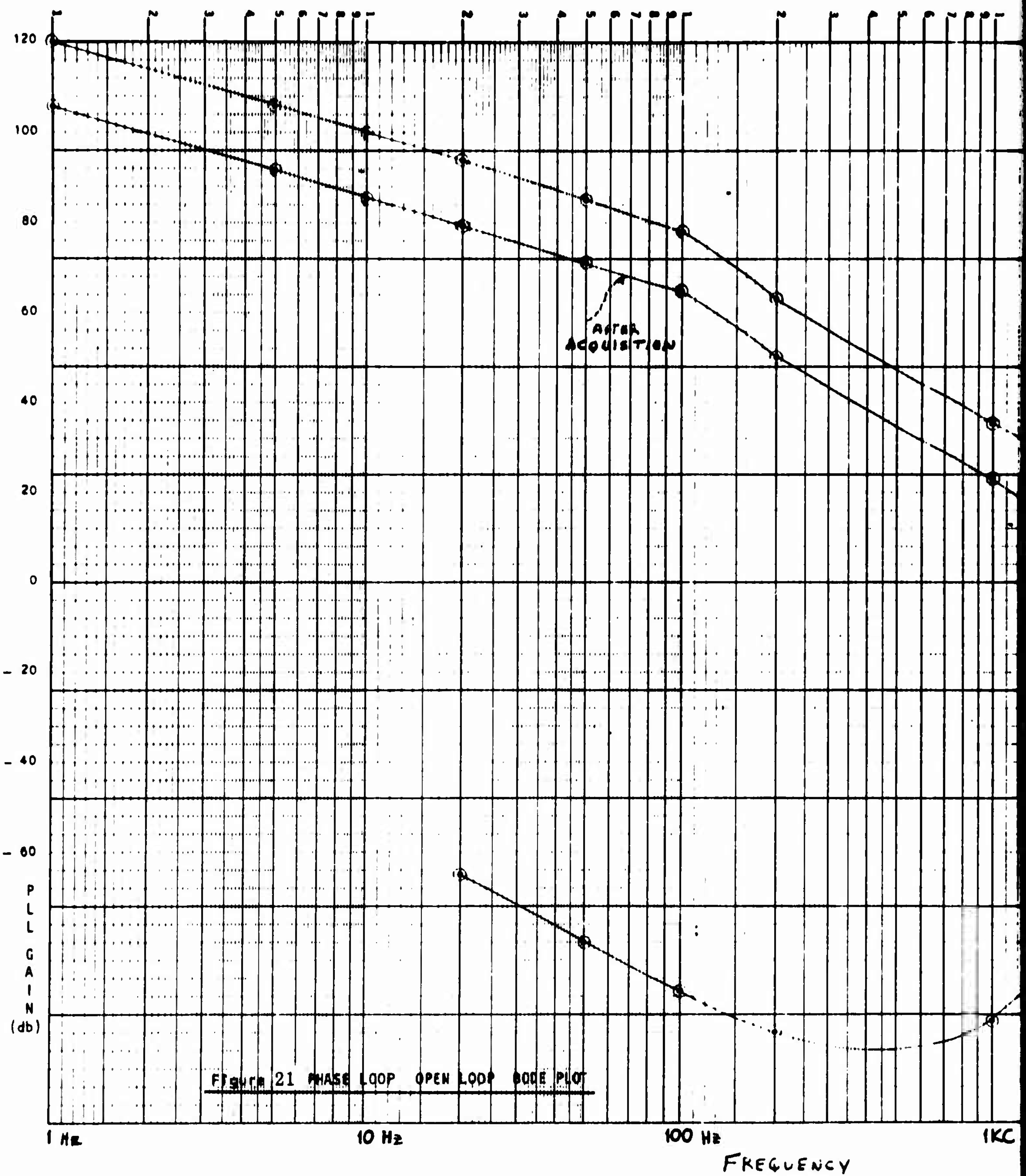
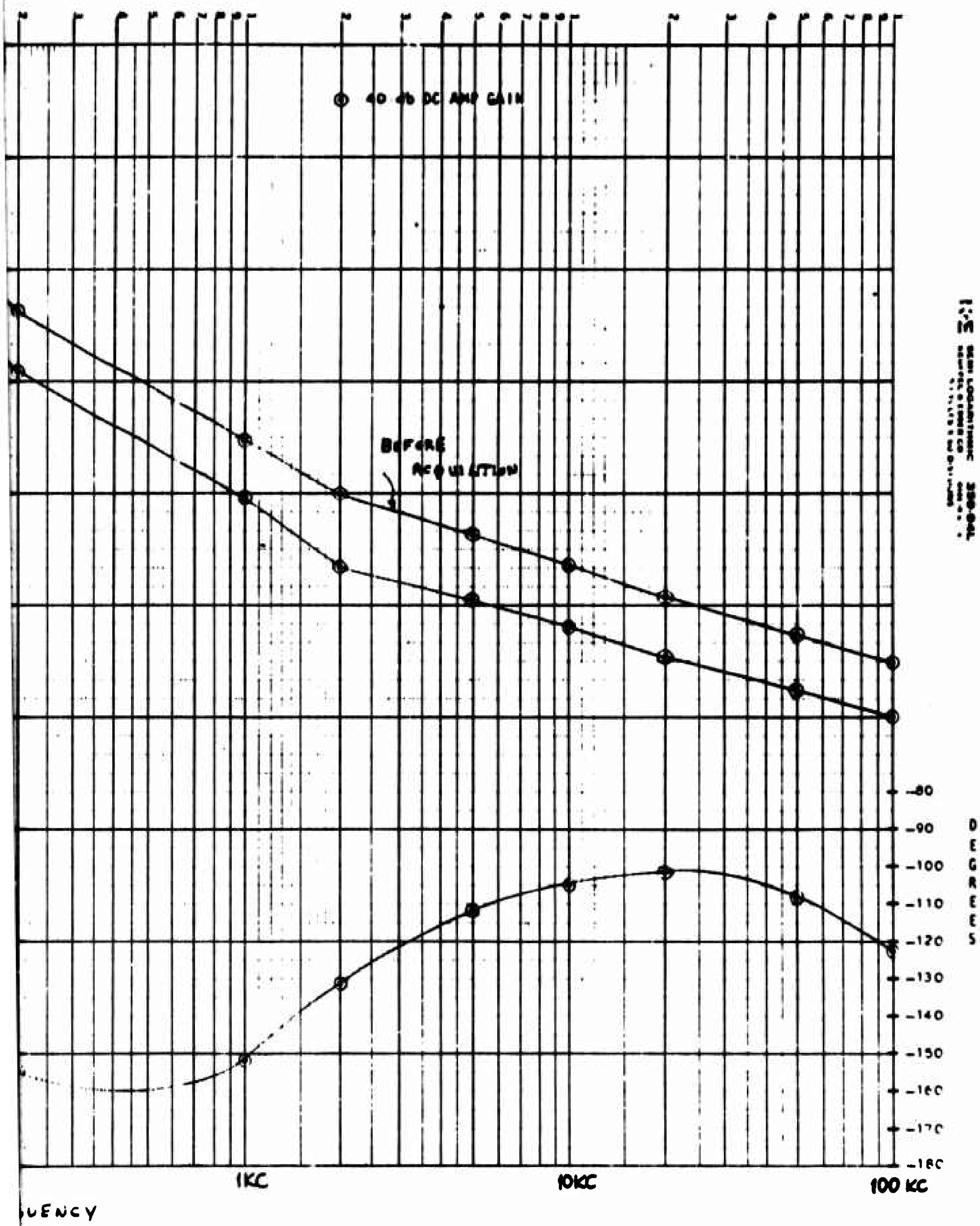


Figure 21 PHASE LOOP OPEN LOOP BODE PLOT

B



40 db DC AMP GAIN
BEFORE ACQUISITION
-80
-90
-100
-110
-120
-130
-140
-150
-160
-170
-180
DEGREES
1KC
10KC
100 KC
UENCY

Figure 21

$KT = -174$ dbm/Hz

$B = 30$ KHz or 45 db

F = Noise figure = 9 db

$S/N = +5$ db before acquisition

S/N after acquisition $= -115 + 174 - 37 - 9$

$= +13$ db

Both figures are well above the threshold of a carrier loop where

$S/N_{\text{Loop threshold}} \simeq +1$ db

Figure 22 shows the MATS PLL acquisition performance locking on a carrier +10 KHz above 420.9375 MHz, to unlock, then 420.9375 MHz -10 KHz, to unlock, and repeat.

$T_{L\ \text{Actual}} \sim 100$ usec.

S/N threshold is measured at $\simeq$ -123 dbm RF signal input to the transponder, (i.e. threshold defined as a 100 Hz unlock to lock rate for a carrier 10 KHz from center frequency).

The slope of the increase in modulation must be such as not to cause the loop to break lock and cause the narrow band crystal filter to ring. The criteria for breaking lock is that the peak phase error of the loop exceeds 90 degrees.

The phase modulation PFFB loop is represented in Figure 23. The filter response is the lowpass equivalent of the bandpass filter. The variable(s) will thus be transformed to a variable about the resonant frequency of the filter.

The transfer function of the loop

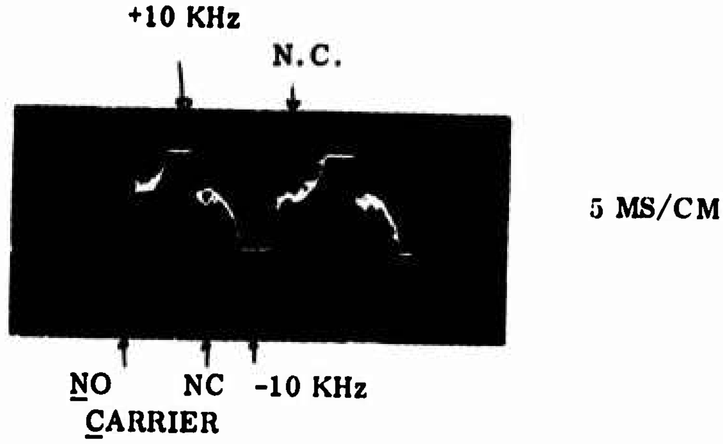


Figure 22. PLL Performance, Dynamic

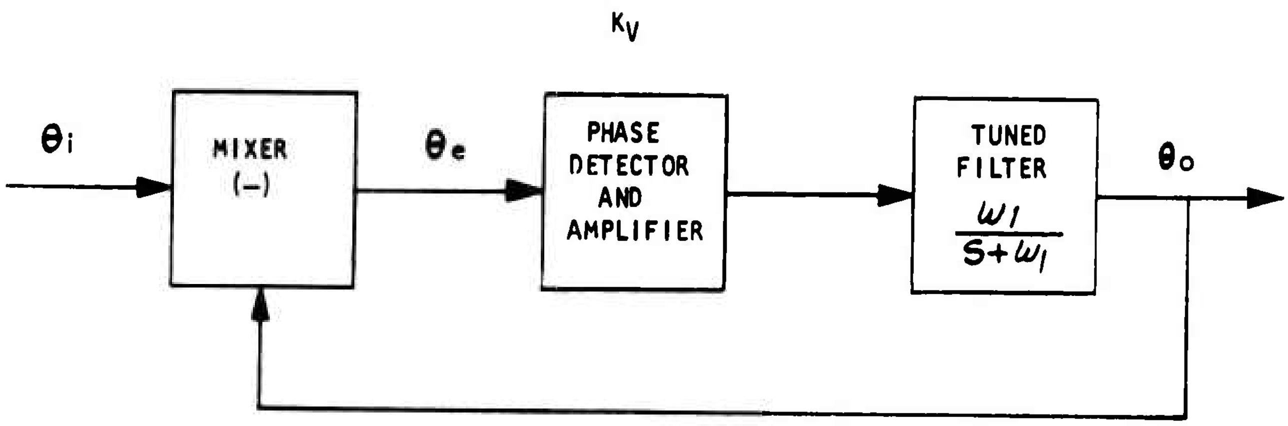


Figure 23. Phase Modulation Compressive Feedback Loop

$$H(S) = \frac{K\omega_1}{S + \omega_1}$$

or

$$\theta_e = \frac{(S + \omega_1)\theta_i}{S + (K+1)\omega_1}$$

Expanding into a Maclaurin's Series,

$$\theta_e = \frac{1}{K+1}\,\theta_i + \frac{K}{(K+1)^2\,\omega_1}\,\dot{\theta}_i \qquad \text{where } \dot{\theta}_i = \frac{d\theta_i}{dT}$$

For the ramp in phase illustrated (Figure 13)

$$\dot{\theta}_i = \frac{2.5 \text{ radians}}{2.5 \times 10^{-3} \text{ seconds}} = 1000 \text{ radians/second}$$

The transponder uses crystal filters having 100 cps 3 db bandwidths. Thus, the lowpass equivalent of $\omega_1 = 2\pi \times 50$ cps and θ_e due to the ramp is equal to (for $K = 30$ db $= 31.6$)

$$= \frac{31.6 \times 1000}{(32.6)^2\, 2\pi \times 50} = 0.094 \text{ radians}$$

However, the overall system will have phase errors introduced due to the ramp of each subcarrier, the frequency error, noise jitter, and the phase modulation of the subcarriers themselves. These errors will be cumulative, as previously explained, and could prevent lockup. The detected signal from the phase detector will be:

$$E_o = k \sin(\theta_{dc} + \theta_1 \cos\rho_1 t + \theta_2 \cos\rho_2 t \ldots + \theta_{ramp}$$

$$\left[f(\rho_1) + f(\rho_2) \ldots + \sigma_n + \ldots \right]$$

As this error approaches 90 degrees, the differential gain for additional error falls to zero and the loop becomes unstable.

Assuming a peak modulation index due to all subcarriers of 15 radians, the error due to the ramp will be 0.56 radians. (15/2.5 x 0.094 = 0.56)

The signal-to-noise ratio in the lowpass loop is a minimum +13 db after acquisition. The signal-to-noise ratio in each subcarrier loop at 2.5 radians of deviation is approximately 20 db. At 0.5 radians, the S/N will be +6 db.

$$S/N_{\text{each subcarrier}} = -KT - S - B - F + M_I$$

where

B = 100 cps x PFFB ratio X Q

= 6220 cps or 38 db

Q = conversion factor from 3 db bandwidth to noise bandwidth

Q = 2 (refer to Appendix C for calculation)

M_I = 2.5 radian or 8 db (20 log 2.5 = 8 db)

F = 9 db

S = 115 dbm

KT = -174 dbm/Hz

$$S/N_{\substack{\text{each subcarrier} \\ \text{2.4 radian}}} = 174 - 115 - 38 - 9 + 8 = +20 \text{ db}$$

One should note that if only one subcarrier is used, but the other 5 are not, then during a S/N measurement

$$S/N_{\substack{\text{measured} \\ \text{2.4 radian}}} = 20 \text{ db} - B_T$$

where

B_T = Noise BW of all subcarrier channels

= 6 B. An 8 db greater noise bandwidth than required for the "one" subcarrier under measurement.

The signal-to-noise ratio during the phase ramp is a variable, being a function of the modulation index. The phase jitter (σ) is a function of S/N.

$$\sigma \text{ rms} = \frac{1}{(S/2N)^{1/2}}$$

The cumulative errors during the ramp will be as follows:

• θ error (ramp)	0.56 radians
• PLL error due to frequency uncertainty	0.1 radians
• Noise Error (rms)	
• Lowpass filter	0.22 radians
• Subcarrier filters	0.10 radians
• θ error at 15 radians peak	$\frac{15}{31.6} = 0.427$ radians

The overall error at -115 dbm is approximately 1.4 radians rms. The system will be stable, if not linear, during the ramp input.

The gain of the PLL loop is sufficient to reduce the phase error (for a frequency offset of 28.5 KHz) to .10 radians. This will require a loop gain of greater than:

$$G = \frac{28.5 \times 10^3 \times 2}{0.1} = 1.8 \times 10^6$$

This is obtained with the following gain constants:

K_ϕ = phase detector gain:	0.1 volts/radian
K_{dc} = dc amplifier gain: (voltage)	46 db
K_{VCO} = voltage controlled oscillator constant:	12 KHz/volt

2.1.18 Grounding (PD para. 3.3.13)

The case of the transponder is isolated from the primary power system. In this regard, DC isolation is $\geq 20\ K\Omega$. The outer case is used only as RF ground.

The primary power is supplied to the transponder power supply using a floating ground system to isolate the transponder case from the primary power system ground. Although isolation at 100 KHz was originally specified at 100 $K\Omega$, capacitance from the primary to case ground was required to reduce RFI problems within the power supply. The infinite DC resistance originally specified is, of course, not practically possible. A DC resistance of 20 $K\Omega$ was found to be a reasonable value for the design.

2.1.19 External Cables (PD para. 3.3.14)

All external cabling other than that utilizing coax, are shielded to the highest degree practical. All shielded cable is grounded (RF ground) at both ends and is of low capacitance. No shield is used as a common return except for RF. Since more than one package is used, provisions are made for a positive electrical RF bond, in a stacked condition, between all package modules.

All cables, external to the modules and carrying RF, are a coax type. The shielded outer conductor is RF grounded on both ends and, when possible, terminated by an impedance equal to the characteristic of the line (i.e., $Z_o = 50\Omega$). Non shield cables occupy the same cable channel as the RF cables. The entire cable channel is RF shielded from the external environment by an aluminum cover plate. The individual modules are machine fitted (to a flatness and parallelism specification) to provide a positive RF bond between them.

2.1.20 Transponder Warm-Up/Shut-Down Properties (PD para. 3.3.15)

The transponder is operating and in a condition to perform as defined in paragraph 3.4 of the purchase description, under the worst combination of environmental, voltage variation and dynamic range conditions defined in PD paragraphs 3.3.6, 3.3.7 and 3.4.1, respectively, within 6 seconds after the receipt of a "select call" subcarrier or the activation of the "select call" override as defined in PD paragraph 3.3.9.

The transponder remains in the "transmit" condition, under the same conditions defined above, at least 7 seconds, but not more than 15 seconds, after the termination of the "select call" subcarrier or the deactivation of the "select call" override as defined in paragraph 3.3.9 of purchase description. This holds true for pulsed "select call" signals where the pulse rate is 20 PPS with carrier on duty cycles of 15 percent or more.

Since the transponder is completely solid state, the original warm-up time of 60 seconds is not required. The effect of this change is to reduce the total energy expended by the satellite power source, during a transmit condition. For example, for three (3) 10-minute transmit periods per day, we have:

Xmit Energy/10 minute period = Xmit power x .167 hrs (i.e. 10 min)

= 39 watts (.167 hrs) = 6.5 watt-hrs

Xmit Energy/3 - 10 min orbits = Xmit Energy/day = 6.5 x 3

= 19.5 watt-hrs

Xmit Energy for warmup = 39 (.05) = 1.95 watt-hrs
(1 MINUTE EACH ORBIT)

Therefore

Total useful energy/day = 19.5 - 1.95 = 17.5 watt-hrs

Since a maximum allowable warm-up of 6 seconds is sufficient, then the total useful energy per day is approximately 19.5 watt-hours versus 17.5 watt-hours/day as was originally intended by the purchase description.

An automatic time delay circuit is incorporated into the power supply module to return the transponder to a "receive" condition upon termination of the "select call" signal, and a "minimum" power drain condition (i.e., "standby") after termination of the coherent carrier.

2.1.21 Telemetry Sensors and Outputs (PD para. 2.1.21)

Certain telemetry parameters are provided within the transponder in order to obtain housekeeping data while in orbit. The circuits required to provide the parameters are integral to the transponder and consist of the following.

2.1.21.1 Thermal Sensors (PD para. 3.3.16.1)

Three (3) thermal sensors are placed in the transponder; one (1) is placed on the case of the power supply; one (1) on a structure in close proximity to the transmitter output; and one (1) on a structure in close proximity to the ranging subcarrier filters and amplifiers. The thermal sensors have a range that covers the expected temperature within the areas of interest when subjected to the Thermal Vacuum test defined in paragraph 4.3.5 of purchase description. (This does not include non-operating storage conditions.) The thermal sensors are of the Fenwal iso-curve type. The sensors have a maximum thermal time constant commensurate with overall system consideration. The isolated leads of the thermal sensors are brought to an external connector on the transponder for use in an external telemetry system. The thermal sensors are passive in that power for these circuits are not provided within the transponder, but from an external source.

2.1.21.2 Input Signal Strength Monitor (PD para. 3.3.16.2)

An active (powered internal to the transponder) circuit is provided within the transponder which is used to indicate the received signal strength at the receiver input of the transponder. The output is in the range of 0.5 to plus 5 VDC, over the dynamic range of -45 dbm to -115 dbm, with external loads of 5.000 ohms or less shunted by 150 micro-micro farads of capacitance. Indications of signal strength are available when the receiver is either in the "receive" or "transmit" condition. All circuits required to provide the above output are power internal to the transponder configuration. The output leads are routed to an external connector on the transponder for use in external telemetry systems. The presence or absence of the above load does not damage the transponder or affect its performance as shown in paragraph 2.2 of this report under the worst combination of environment, voltage variation and dynamic range. The input strength output is approximately logarithmic. At the given temperature, the resolution of the output is ± 2 db. The output has an overall reading accuracy, including resolution and stability, over the worst combination of environment conditions and dynamic range of ±5 db for a period of one year.

Since the transponder contains two types of AGC (1) non-coherent (signal plus noise)AGC and (2) coherent (signal)AGC, the signal strength indication over the -45 dbm to -115 dbm range is shared between two outputs. The range -45 to -105 dbm is covered by the signal plus noise AGC. The remainder is covered by the signal AGC. The signal AGC is fed to telemetry on the same line as the "frequency acquisition" indicator. In the interest of conserving power to the maximum extent possible in the "standby mode," the AGC circuitry is not turned on until an inband carrier provides a correlated output from the detector, placing the transponder in the "receive" mode.

The original specification uses words "approximately logarithmic" to define the signal strength monitor characteristic. This should be redefined with reference to the actual curves obtained, since the words are too subjective to accurately denote the output characteristic.

2.1.21.3 Power Output Monitor (PD para. 2.1.21.3)

An active (powered internal to the transponder) circuit is provided to indicate the power of the 449.00 MHz transmitter output. This circuit is capable of providing outputs within an 0 to plus 5 volts DC range for either of the 1.5, 3.5 or 4.5 watt output configurations. The output voltage of this circuit versus the output power of the transponder is stable within ±10% of maximum reading and approximately linear over all conditions of environment and voltage variations defined in this document.

This circuit is capable of driving 5000 ohms shunted in parallel with 150 micro-micro farads. When the transponder is in a "standby" condition, the voltage out of this circuit is "0" volts. All circuits and power required to provide the output described above, are an integral part of the transponder. The output leads are routed to an external connector on the transponder for use in external telemetry systems. The presence or absence of the above load does not damage the transponder or affect its performance as called out in paragraph 2.2 of this report under the worst combinations of environment, voltage variation or dynamic range.

The original output linearity specification of 5% is too tight for simple RF diode detection circuitry. Unless one can justify the additional circuitry required to linearize this function, it is reasoned that the present circuitry is adequate.

The plus 0.5 volts originally required during "standby" is not provided, since no power is supplied to the transmitter module during standby.

2.1.21.4 Automatic Phase Control and Frequency Acquisition Voltage (PD para. 3.3.16.5)

Two outputs, indicative of Automatic Phase Control and of Frequency Acquisition, are provided. The output impedance of these circuits is $\leq$ 600 ohms. These circuits are capable of driving loads of 5000 ohms or less, shunted by 200 microfarads capacitance at levels between 0.5 and 5.0 volts for the output range. Each output is plotted for each transponder.

The APC monitors the error voltage of the phase lock loop through a buffer amplifier whose output impedance is $\cong$ 150 ohms. The Frequency Acquisition output monitors the correlation detector output which also indicates coherent AGC voltage. The output impedance is $\cong$500 ohms.

2.1.22 Transponder Detector Output (PD para. 3.3.17)

An output is provided from the output of the transponder detector. This output is used to drive command circuits external to the transponder. Access to this circuit is available at an external connector. The output impedance of this circuit is 600 ohms or less. This circuit is capable of driving a load of 5,000 ohms or less shunted by 200 micro-micro farads capacitance at a level of at least 0.5 volts RMS. Insertion or removal of a load as defined above does not damage the transponder or degrade its performance as called out in paragraph 3.4 under any worst combination of environment, primary voltage variations and dynamic range.

The phase detector output is fed to a wideband video amplifier providing a level of at least 1$\sqrt{}$ PP of composite signals, including wideband noise. Individual signal amplitudes are a function of their modulation index. The noise level is a function of the carrier signal level into the transponder.

2.1.23 **Command Frequency Outputs** (PD para. 3.3.19)

Two output circuits are provided within the transponder which are used for command signal outputs. The circuits operate on subcarriers within the range of 400 - 600 KHz and utilize crystal filters in the same manner as the select call circuit. The output impedance of these circuits is 600 ohms or less and capable of driving loads of 5000 ohms or less shunted by 150 micro-farads at levels of at least 1.0 volts RMS. Bandwidth of the filter circuits is 100 cps. They provide a S/N improvement of 40 db or a total S/N of at least 19 db with input signal levels of -115 dbm.

The command signal is not part of the loop. Detection bandwidth is the band width of the crystal filter. The modulation index for the command subcarrier is 0.5 radian.

The S/N is given by

$$S/N = -KT - 115 \text{ dbm} - B - F \cdot \Theta_m$$

where

$$B = 100 \text{ cps} = 20 \text{ db}$$

$$\Theta_m = 0.5 \text{ radians} = -6 \text{ db}$$

$$F = 9 \text{ db}$$

$$S/N = 174 - 115 - 20 - 9 - 6 \text{ db} = 24 \text{ db}$$

$$S/N = 24 \text{ db}$$

Thus the +19 db is easily met. The crystal filters are multiple pole types where the 3 db bandwidth is a good approximation to actual noise bandwidths. The mod.index can be anywhere within 0.3 to 0.5 radians. Higher indexes can be used if all other subcarriers are not on simultaneously. A buffer amplifier provides a low impedance drive of 1$\sqrt{}$RMS at $\cong$150 ohms to the telemetry.

2.1.24 Limiting (PD para. 3.3.21)

Positive transistor limiting is included both in the receiver and transmitter sections of the transponder. Transistor amplifiers whose characteristics are such as to provide symmetric limiting(cutoff and saturation points) are used in place of the more common diode limiters. These limiters are convenient and provide equal charge and discharge paths for both positive and negative voltage clipping.

2.2 Performance (PD para. 3.4)

The transponder performs in the manner called out below under the worst combination of environment, primary input voltage variations, and dynamic range.

2.2.1 Dynamic Range (PD para. 3.4.1)

The transponder performs and operates as per specification (Section V) over input power ranges from -45 dbm to -115 dbm with any combination of fixed frequency subcarriers defined in paragraph 3.1.1 of the purchase description.

As mentioned previously, both signal plus noise and signal AGC circuitry are used to cover the complete dynamic range. The signal plus noise AGC is of the reverse type in order to provide a large control range. Under usual circumstances, a design based upon reverse AGC would suffer from poor dynamic range and large variations in transistor parameters, but since the IF strip uses integrated circuits, where multiple active circuits are contained within each unit, the old transistor "cliches" are not necessarily valid. For the circuits chosen, reverse AGC doesn't adversely affect either the transistor operating points, dynamic range, linearity or input-output impedance, but does provide a constant signal output over wide variations in input level. The gain control per stage is controlled from cutoff to maximum gain. The detection bandwidth is 2 MHz, thus providing AGC control from -45 dbm to -100 dbm.

The coherent AGC (Signal type) is a forward type which utilizes a correlation detector to detect the signal, followed by a dc amplifier, whose output is proportional to the transponder carrier signal input, and a single transistor IF amplifier as the controlled element. Since the bandwidth of the correlation loop is approximately the same as the PLL, a S/N ratio of = +13 db for -115 dbm signal level is expected. The coherent AGC provides +15 db of control after the transistor limiter circuit, but before the phase detector. Thus, the carrier signal level into the PLL is held constant independent of the carrier signal levels into the transponder, to below -115 dbm.

2.2.2 Noise Figure (PD para. 3.4.2)

The noise figure of the transponder receiver is 6.7 db.

The noise figure of the receiver is given by the relationship:

$$F_T = \frac{T_A}{T_O} + F_1 - 1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \frac{F_4 - 1}{G_1 G_2 G_3} + \frac{F_5 - 1}{G_1 G_2 G_3 G_4} + \frac{F_6 - 1}{G_1 G_2 G_3 G_4 G_5}$$

where F_1 and G_1 are related to the losses in the preselector, F_2, F_3 and G_2, G_3 are the noise figures and gains of the RF stages, respectively. F_4 is the noise figure of the 2-pole filter, F_5 is the mixer noise figure, F_6 is the IF noise figure, and G_5 is the mixer insertion loss. T_A is the antenna temperature and T_O is 300° Kelvin. Figure 24 is a block diagram of the receiver front end.

The selectivity of the preselector is kept at the minimum level required to prevent desensitivation of the RF amplifier. This is done to reduce its insertion loss. An eight-pole filter is used since a bandwidth of 18 MHz and an average 65 db rejection of the 449 MHz signals is required. A grounded base amplifier is used as a first and second RF amplifier. Additional RF selectivity requirements are placed in the collector of the RF amplifiers where its effect on noise figure is minimal.

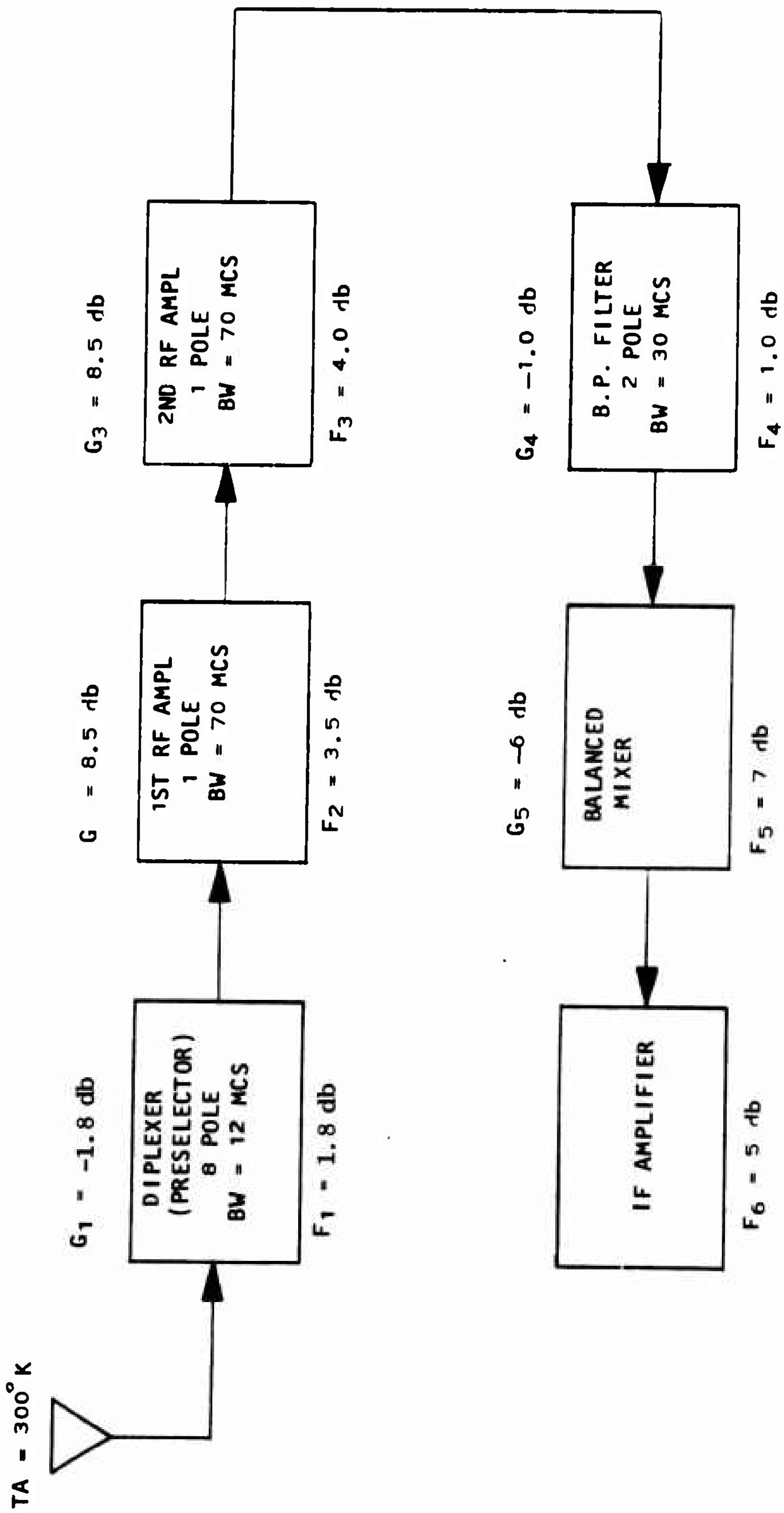


Figure 24. Multi-Altitude Transponder Receiver RF Noise Figure Block Diagram

The values of the parameters are:

$T_A = 300^oK$

$T_O = 300^oK$

$F_1 = 1.8 \text{ db} = 1.56$

$G_1 = 0.64$

$G_2 = 8.5 \text{ db} = 7.2$

$G_3 = 8.5 \text{ db} = 7.2$

$F_3 = 4.0 \text{ db} = 2.5$

$F_2 = 3.5 \text{ db} = 2.2$

$G_4 = -1.0 \text{ db} = 0.82$

$G_5 = -6 \text{ db} = .25$

$F_4 = 1.0 \text{ db} = 1.2$

$F_5 = 7.0 \text{ db} = 5$

$F_6 = 5 \text{ db} = 3.2$

$$F_T = \frac{300}{300} + 1.56 - 1 + \frac{2.2 - 1}{0.64} + \frac{2.5 - 1}{(0.64)\,(7.2)} + \frac{1.2 - 1}{(0.64)\,(7.2)^2} + \frac{5 - 1}{(0.64)\,(7.2)^2\,(0.82)} + \frac{3.2 - 1}{(0.64)\,(7.2)^2\,(0.82)\,(0.25)}$$

$$F_T = 1.56 + 1.88 + 0.325 + 0.0063 + 0.156 + 0.344 = 4.27 \text{ or } 6.3 \text{ db}$$

Note that the primary contributions to the the overall noise figure are preselector and the first RF stage. Note that the design of all circuits, as far down as the IF strip, exhibit excellent noise performance denoting the care which was taken to provide the minimum noise figure consistent with other specification requirements.

2.2.3 Phase Stability (PD para. 3.4.3)

The phase delay imparted to any ranging or timing subcarrier in the presence of any combination of subcarriers as defined in paragraph 3.1.1 of purchase description and at individual modulation index as defined in purchase description 3.3.12 does not change from the phase referenced at a -75 dbm signal input to the transponder receiver, over the dynamic range of -45 dbm to -115 dbm by more than that shown below. (NOTE: Also referenced to room temperature.)

Temperature Range	Phase Stability Over Dynamic Range
Minus 4°F to Plus 160°F	2.5 degrees maximum
Plus 23°F to Plus 113°F	0.75 degrees maximum
Plus 32°F to Plus 95°F	0.5 degrees maximum

Once set at -45 dbm and -115 dbm, the phase delay imparted to any subcarrier, under the conditions as defined above, remains constant with 1.0 degree for periods up to 45 minutes. Hysteresis effects are constant to within 0.1 degree for both the conditions of start-up/shut-down and changes of input signal over the dynamic range. The above holds true for incoming signals having AM components up to 35 percent.

The phase stability systems analysis was thoroughly covered in paragraph 3.1.3 of this report. The basic circuit design technique is to develop a good open loop transponder to ensure excellent closed loop stability. For all circuit designs, consistent with other specifications, the phase delay stability is of primary importance. Circuits are generally of broadband design to reduce their absolute phase delay. The primary factor, that is the crystal filters, are, as mentioned previously, carefully chosen. The ranging and timing crystal filter bandwidths are kept at the widest bandwidth possible, to reduce their phase slope, consistent with the S/N restrictions.

Circuitry in the feedback loop is minimized to a single 2-pole cavity filter to reduce the effects of feedback delay changes on output phase shift.

Circuitry outside the PFFB loop is kept as wide as possible, again to reduce absolute delay times.

In general, for most of the circuit design, a large power mismatch was purposely incorporated into the design to reduce the effects of active component parameter shifts on the circuit transfer function. Thus, for example, the Q of a tuned circuit was determined by a discrete loading resistor rather than the input-output impedance of the associated active devices.

In referencing the original specifications of paragraph 3.4.3 in the purchase description, it is interesting to point out that as written, the only phase variations of interest are those over dynamic range -45 to -115 dbm, referenced to -75 dbm, the temperature range specified. This, we are sure, is not the intent of the customer and thus, it is suggested that a reference temperature be stated.

Example:

Original Specification – $\pm 3^{o}$ Maximum for dynamic range -45 to -145 dbm referenced to -75 dbm at any temperature $-4^{o}F$ to $+160^{o}F$

Modified Specification – $\pm 3^{o}$ Maximum for dynamic range and temperature specified referenced to -75 dbm at room temperature

The phase stabilities originally specified can be changed to those indicated above, namely

$\pm 3°$ Maximum to 2.5° Total

$\pm 1.5°$ Maximum to 0.75° Total

$\pm 1.0°$ Maximum to 0.5° Total

Hysteresis can be specified as 0.1°. The above changes are practical since the design of the transponder is such as to allow strict control of the phase stabilities with temperature and dynamic range beyond the original specification limits of the purchase description.

Some additional causes of phase delay variations not previously mentioned are:

- Amplitude modulation in the transmitter output at the subcarrier frequencies. This AM can be caused by the phase modulation process and by PM-to-AM conversion in the transmitter amplifiers. The amount of AM is reduced by using broadband transmitter amplifiers, careful design of the phase modulator, and by limiting in the receiver and in the transmitter chain.
- The PFFB loop will not compensate for variations in time delay brought about by the AM components fed back in the loop. Further, the AM is magnified by the reduction of the PM as a result of the feedback.
- Spurious signals in the local oscillator path. Balanced mixing is incorporated to reduce feedback of spurious signals at the receiver frequencies from the LO path.
- Leakage LO paths. The design of the diplexer and resultant RF selectivity keep this leakage down to at least -30 db.

2.2.4 <u>Select Call Sensitivity</u> (PD para. 3.4.4)

The "select call" circuit positively engages at levels of -115 dbm or less at the selected frequency. "Positively engage" is interpreted to mean continuous operation without interruption. The above holds true for incoming signals having AM components up to 35%.

The original specification modification was discussed in paragraph 2.1.14 of this report. Note that for the initial select call operation, only the "select call" subcarrier, in combination with, say, a command subcarrier, should be transmitted by the ground station. Once the initial select call operation is accomplished, all combinations of composite input signals and modulation levels can be used simultaneously to keep the transponder in the "transmit" condition.

2.2.5 <u>Transmitter Modulation</u> (PD para. 3.4.5)

The transmitter employs phase modulation techniques. The modulation index of each subcarrier re-transmitted at the 449.00 MHz carrier lies within the range of 0.5 to 2.5 radians, and the re-transmitted modulation index for each subcarrier is the same -3 to -23% as each incoming subcarrier signal modulation index. The modulation index of each subcarrier re-transmitted at the 224.500 MHz carrier lies within the range of 0.25 to 1.25 radians and the re-transmitted modulation index for each subcarrier is one-half of that re-transmitted at 449 MHz. This shall hold true for any combination of incoming signals as defined in paragraph 3.1.1 of purchase description whose modulation indexes for each subcarrier are within the limits of 0.5 to 2.5 radians. The above shall hold true for incoming signals having AM components up to 35%.

Since the 224.5 MHz transmitted signal is derived from the same multiplier chain as the 449 MHz transmitted signal, then the modulation index of each transmitted subcarrier at 224.5 equals one-half of that transmitted at 449 MHz.

The original purchase description specification stated that the modulation index at the transponder transmitter output has to be within ±10% of that received. What is the theoretical best that can be done using a PFFB type of transponder.

$$m_o = m_i - \frac{m_i}{1 + K\mu\beta}$$

where

m_o = output index

m_i = input index

$k\mu\beta$ = system loop gain

$$\frac{m_i}{1 + K\mu\beta} = \text{IF index}$$

for

$$K\mu\beta = 30$$

$$m_o = m_i - \frac{m_i}{31}$$

$$m_o = m_i \left(1 - \frac{1}{31}\right) = m_i \quad (.968)$$

therefore, the output modulation index is 96.8% of the input modulation index, at its theoretical best. Allowing the same variation of 20% as the original specification, we have

$$m_o = \left.\begin{matrix} -3\% \\ \\ -23\% \end{matrix}\right\} m_i$$

2.2.6 Transmitter Output Spurious Suppression (PD para. 3.4.6)

The spurious emission at the transmitter is in accordance with paragraph 3.5.2 of MIL-I-11748B, as amended.

Transmitter spurious is reduced primarily by the post filter at the 224.5 and 449 MHz output. Since the transmitted carriers are derived from a fundamental 18.7 MHz oscillator, the multiplication up to the output frequencies is obtained in small steps rather than a single jump. This is done in order to filter the fundamental to guarantee the output rejection specifications of $\geq$ 60 db for all harmonically related spurious. 18.7 MHz is first multiplied by 4 to 75 MHz. A four-pole filter at 75 MHz rejects 18.7 MHz by 50 db minimum. A times three multiplier yields 224.5 MHz followed by a times 2 for the 449 MHz. Four poles at 224.5 reject the 75 MHz by > 70 db (two of these poles are the 224.5 MHz post filter). Thus the 18.7 MHz suprious is rejected an additional 24 db above and beyond the 50 db obtained by the 75 MHz filter.

The 8 pole 449 MHz post filter rejects 75 MHz > 100 db while the 18.7 MHz spurious is rejected by > 50 db above and beyond that obtained by the filtering at 75 MHz.

2.2.7 Receiver Spurious Response (PD para. 3.4.7)

The response of the transponder receiver to spurious signals is down at least 60 db. This includes response to image frequencies, intermediate frequencies and unwanted signals generated within the receiver. The above holds true for incoming signals having AM components up to 35%.

The image frequency is = 2 IF + RF = 2 (28 MHz) + 421 MHz = 477 MHz. The 8-pole preselector provides 100 db of image rejection without considering the other poles provided by the preamplifier circuitry. Rejection of unwanted signals received through the transponder antenna is excellent due to this 8-pole

preselector, the use of double conversion, and the use of selective IF amplifiers rather than wideband RC or DC coupled amplifiers having good gain characteristics at the IF frequency.

Frequencies generated within the transponder are confined by extensive line filtering between modules, and individual circuits. In addition, all oscillators within the transponder are RF shielded by enclosures and use shielded coax cable to deliver their outputs.

2.2.8 Amplitude Modulation Suppression (PD para. 3.4.8)

The transponder receiver provides sufficient limiting to suppress all AM components on the receiver carrier to less than 5 percent as measured at the output of the receiver detector with any combination of input signals as defined in paragraph 3.1.1 of purchase description. The output of the transponder transmitter has no AM components of more than 5%. The above holds true with signals at the receiver input having AM components up to 35%.

Some of the most severe problems encountered in a phase-following system are those resulting from the presence of amplitude modulation components on the received signal, or the generation of such components internal to the transponder. Since the phase detector is sensitive to amplitude variations, these amplitude modulation signals are faithfully reproduced and transmitted to the frequency modulator where they are introduced into the frequency following loop as FM signals. The transponder now attempts to make a frequency modulation correction to an amplitude modulation signal, with the result that these undesired AM effects create distortion and are re-transmitted to the ground station as a part of the FM composite signal. Since these amplitude modulation products are not desired, the MATS transponder use three separate circuit functions to eliminate them. The AM either transmitted or inherently generated is reduced by (1) a limiter in the receiver, (2) a limiter in the transmitter and (3) the receiver

AGC system. The receiver and transmitter limiters reduce the AM by suppression of all composite signal amplitudes to a fixed value. Thus, the output amplitude is relatively independent of the input amplitude variations and represents a constant output power level.

The receiver non-coherent AGC can follow AM signals whose frequency response is $\leq$ 10 KHz. Since the AM transmitted by the ground station occurs primarily at 256 Hz then the non-coherent AGC reduces this AM commensurate with its feedback gain. Refer to Figure 25 for non-coherent AGC time response.

The coherent AGC will also react, in a similar manner, to the low frequency AM resulting from the residual AM fed to the receiver correlation detector. Refer to Figure 26 for coherent AGC time response.

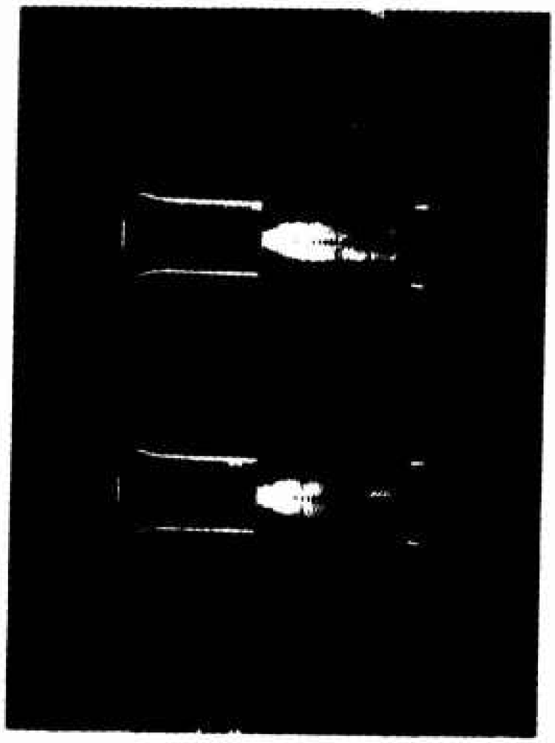
Note that when the ground transmitter has AM on its carrier, the effective carrier power is reduced when compared to the same signal level without AM. For 35% sine wave AM the average carrier power reduction is approximately 2 db. This should be considered as an additional factor when calculating the MATS transponder receiver sensitivities since they are measured with 35% AM at the receiver input.

2.2.9 Data Transit (PD para. 3.4.9)

The rise time of any combination of composite data signal, as defined in paragraph 3.1.1 of purchase description when pulsed on and off at an 80 PPS or less rate with a 10 plus 1 minus zero millisecond "carrier on" duty cycle at the transponder receiver input, does not exceed 1 millisecond or have an overshoot of more than 5% as measured at the output of the receiver detector (filtered output). The fall time of the composite data, under the same conditions, does not overlap any succeeding pulse by more than one millisecond. Undershoot does not exceed 5%.

The data transit time is controlled primarily by the 3 db bandwidth of the PFFB loop. The bandwidth is:

NOT REPRODUCIBLE



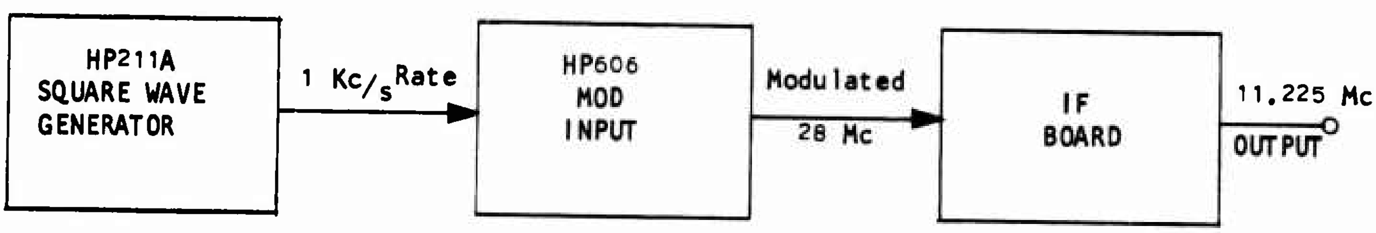


Figure 25. AGC Time Response (Noncoherent Loop)

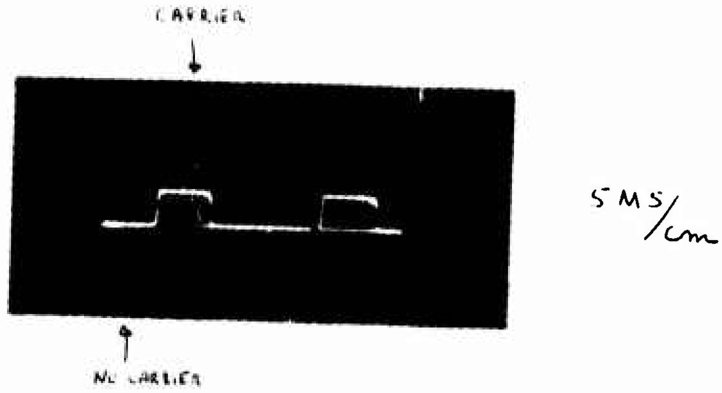


Figure 26. Coherent AGC, Dynamics

$$Bw_{3\,db} = (3\text{ db crystal filter bandwidth}) \times \text{PFFB ratio}$$

$$= 100\text{ cps} \times 30 = 3\text{ KHz}$$

$$tr = \frac{0.35}{BW_{3db/2}} = \text{rise time } (10 - 90\%) \text{ for a step input}$$

$$t_r = \frac{.35}{1.5\text{ KHz}} = .234\text{ MS}$$

Refer to Figure 27 for actual measured times. Note that the noise output, without a subcarrier, was measured since the dynamics required of a subcarrier for the above test is not conveniently obtained.

Since the loop is, if anything, overdamped, overshoot problems are minimized.

2.2.10 Frequency (PD para. 3.4.10)

The output frequency of the transponder transmitter is 224.500 MHz and 449.000 MHz. The accuracies and stabilities of the 449 and 224.5 MHz output are ±0.001% or better over the worst environmental conditions defined in this purchase description.

The frequency accuracy and stabilities of the transponder transmitter frequencies are determined by the 18.7 MHz oscillator from which the output frequencies are derived. The oscillator is preset by a trimmer capacitor to within ±1 Hz of the desired output frequencies 224.500 MHz and 449.000 MHz (when multiplied by 12 to 224.5 and by 24 to 449 MHz). The stability of the oscillator is specified to better than ±.001% over the worst environmental conditions as specified in the purchase description.

2.2.11 Ranging Sensitivity (PD para. 3.4.11)

The ranging sensitivity of the transponder is at least -115 dbm. Ranging sensitivity is defined as that signal level at the receiver input, with any combination of the modulation subcarriers defined in paragraph 3.1.1 of purchase description at which each of the ranging or timing subcarriers, as defined in purchase description subparagraphs a, b, c, d, e or f has a signal-to-noise ratio of at least 6 db at the input to the transmitter modulator.

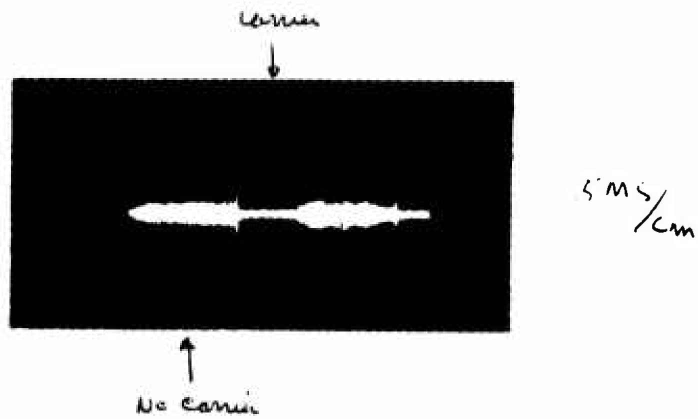
Much has already been said relative to the above in purchase description paragraph 3.3.12. The main point that should be made here is that the value of signal-to-noise ratio which can be easily measured must be manipulated to approach the above 6 db figure. Therefore, it is recommended that the above specification be rewritten to encompass the result of the actual measured value.

Re-write to: "Ranging sensitivity is defined as that signal level at the receiver input, with one ranging or timing subcarrier at a modulated index of 2.4 radians which produces a signal-to-noise ratio of at least 12 db at the input to the transmitter modulator."

2.2.12 Transmitter Carrier Signal-to-Noise Ratio (PD para. 3.4.12)

The transponder transmitter carriers (449.000 MHz and 224.500 MHz) in the 1.5, 3.5 and 4.5 watt configuration, have a signal-to-noise ratio of at least 40 db. The above holds true when the transmitter output has been modulated by each of the subcarriers defined in subparagraphs a, b, c, d and e of paragraph 3.1.1 in the purchase description at the level defined in paragraph 3.4.5.

There has been a long misunderstood specification, the objective of which is to guarantee that the transmitted output of the transponder does not degrade that S/N ratio presented by the receiver to the modulator. This is easily guaranteed by providing a carrier to noise ratio of at least 40 db at the output.



NOT REPRODUCIBLE

Figure 27. PFFB Loop Dynamics

Another test would be to compare the S/N output of the transmitter with that presented to the modulator input.

The S/N ratio of the transmitter alone is calculated in Appendix B and results in a minimum S/N of 60 db.

SECTION II

COMPONENT SELECTION CRITERIA

2.0 CONSIDERATIONS FOR COMPONENT SELECTION

2.1 Selection of components was made after a careful review of design requirements for the part, and wherever possible, its predicted reliability. Table V lists the component parts of the transponder, their description, and the reasons for selection.

TABLE V

List of Component Parts

Part No. & Description	Criteria of Selection
MC-1110 (Motorola) Linear Integrated Circuit	Emitter-coupled, integrated circuit linear amplifier for IF and RF amplications. Frequency range is DC to 300 MHZ. This amplifier is used primarily in the transponder IF circuitry since it exhibits excellent AGC characteristics, high port impedances, and symmetrical limiting.
CK Type Ceramic Capacitor	A ceramic capacitor conforming to MIL-C-11015C/18 and /19. A highly reliable component exhibiting good performance as a bypass and coupling capacitor thru the VHF frequency band.

Part No. & Description	Criteria of Selection
CM Type (Arco) Mica Capacitor	A dipped mica capacitor conforming to MIL-C-5/18. These capacitors are used primarily as tuned elements in filter circuits associated with both passive and active networks. They are chosen primarily for their high Q performance up through UHF. Other important features are stable electrical characteristics, especially with temperature, large value selection, small size and excellent reliability.
CS Type (Arco) Tantalum Capacitor	A solid electrolyte tantalum capacitor conforming to MIL-C-26655A. This capacitor is used primarily as a low frequency-to-MF bypass. Its use is desirable due to its large capacity-to-volume ratio.
JMC 2951 (Johanson) Variable Capacitor, Air	A subminiature variable air capacitor used in all tuned filters in the VHF range. This capacitor is chosen primarily for its exceptional high Q (>1000) over a large (10:1) capacitance tuning and frequency range. Other important features are stable electrical characteristics, high breakdown voltage, and small size. Some installation problems have been encountered since the capacitor top is fastened with a solder connection. All capacitors are ordered with high temperature solder and installed with extreme care.
SS5A (Allan Bradley) Bypass Capacitor, High Frequency	A ceramic button capacitor for VHF-UHF applications. Used primarily as an RF bypass capacitor because of its low inductance, high capacitance, and small size.
2E1000RK (Mucon) Coupling Capacitor, High Frequency	A ribbon lead capacitor used, and specially designed for, UHF applications as a coupling and RF bypass capacitor.

Part No. & Description	Criteria of Selection
2934-000 (Erie) Standoff Capacitor	A button mica standoff capacitor conforming to MIL-C-10950. This capacitor is used as a UHF bypass and coupling in the transponder transmitter. Their high Q and low inductance make them ideal for this application.
301-000 Type (Erie) Ceramic Capacitor	A tubular ceramic capacitor chosen primarily for its temperature compensating ability, stability, high Q, close tolerance and value selection.
TRW V900EA Varicap, Variable Capacitor	This varicap is a silicon alloy P-N junction device designed for use as a voltage variable capacitor. The transponder modulator uses this particular type due to its high value (100 pf) of nominal capacitance, excellent tracking performance, low leakage and high Q.
HPA 0182/3 Step Recovery Diode	This diode is used for high order, single stage, harmonic generation. It conforms to MIL-S-19500 requirements for application in military equipment.
1025/2500 (Delevan) Inductor	This RF coil is used as a tuned element in various filter circuits due to its small size, large value selection, high resonant frequency, high Q and low DC resistance. It is also used as a line filter element for all individual circuits in the transponder. Some reliability problems have been encountered with this component. An extensive investigation is presently taking place.
2N3287 Transistor	A high frequency transistor chosen primarily for its excellent forward AGC characteristics.
2N3137 Transistor	A high frequency transistor chosen primarily for its large signal capability in the VHF-UHF region.

Part No. & Description	Criteria of Selection
2N918 Transistor	A high frequency transistor for use as a general purpose amplifier from DC to UHF due to its excellent gain bandwidth product, low noise figure, and complete manufacturers documentation of performance and reliability.
RC Type Fixed Composition Resistor	Fixed composition resistors are used throughout the transponder due to their small size and high reliability. Values above 4.7 meg are not used due primarily to possible serious radiation effects.
TM 1/4 (Texas Instruments) Sensistor	A positive temperature coefficient transistor for use as a resistive temperature compensating element. This component exhibits a large positive coefficient making it an easily adaptable general purpose compensating element when combined with fixed resistance values. This component is designed to pass military test conditions as per MIL-S-202.
2N3880 Transistor	A UHF silicon transistor chosen for its exceptionally low noise figure at the transponder received frequency.
1N831 Diode, RF	An RF diode used as an RF-AM detector. Chosen primarily for its RF detection efficiency.
2N4040 Power Transistor, RF	A high power VHF transistor used as a final RF power output stage (224.5 MHZ) in the transponder transmitter. Chosen for its "State of the Art" efficiency.
3TE440 Power Transistor, RF	A high power UHF transistor used as a final RF power output stage (449 MHZ) in the transponder transmitter. Chosen for its "State of the Art" efficiency.

Part No. & Description	Criteria of Selection
μA702A Integrated Circuit, Linear	The μA702A is a high gain, wideband DC amplifier with differential inputs providing useful feedback gain from DC to 30 MHZ. The choice of the μA702A was based primarily on its differential operational performance at very low primary input power levels. Since primary input power requirements are critical in the transponder design, this amplifier is used extensively in all baseband and telemetry buffer circuits.
1N702A Zener Diode	A silicon zener diode used primarily in series voltage drop applications. Chosen primarily for its low zener voltage and close tolerance at small operating currents.
1N914 Diode	A high speed general purpose silicon diode for use in assorted transponder circuits. Chosen primarily for its high back resistance, high forward conductance and speed.
JFD LCS LC Tuner	An LC tuner conforming to MIL-C-14409B. A compact component combining a fixed ribbon inductor and a variable air capacitor. This component is chosen primarily for its high Q, stable characteristics, and small size.
05395 (N Seves Kemet) Tantalum Capacitor	A non-polar solid tantalum capacitor chosen primarily for its small size.
SC1651G Diode Quad.	A special diode arrangement for use in the phase detector circuits of the transponder receiver. Chosen for its closely matched diode characteristics.
3TE450 Power Transistor, RF	A medium power UHF transistor used as a driver stage in the transponder transmitter. Chosen for its "State of the Art" efficiency.

Part No. & Description	Criteria of Selection
3TX132 Power Transistor, RF	A low-to-medium power VHF transistor used as a driver in the transponder transmitter. Chosen for its "State of the Art" efficiency.
2N2222 Silicon Transistor	A high frequency silicon transistor for use in the transponder power supply. Chosen primarily for its high DC beta.
2N3250 Silicon Transistor	A high frequency silicon transistor used as a high speed switch in the transponder power supply. Chosen primarily for its high speed, low stage, low offsets and complete published parameters.
2N3853 Silicon Transistor	A silicon power transistor used in the power supply. Chosen primarily for its high speed on power switching capability and low offsets.
1N4572 Zener Diode	A low-level temperature compensated zener reference diode. Chosen primarily for its reliable operation at low current levels.

APPENDIX A

THE POST FILTER PROBLEM

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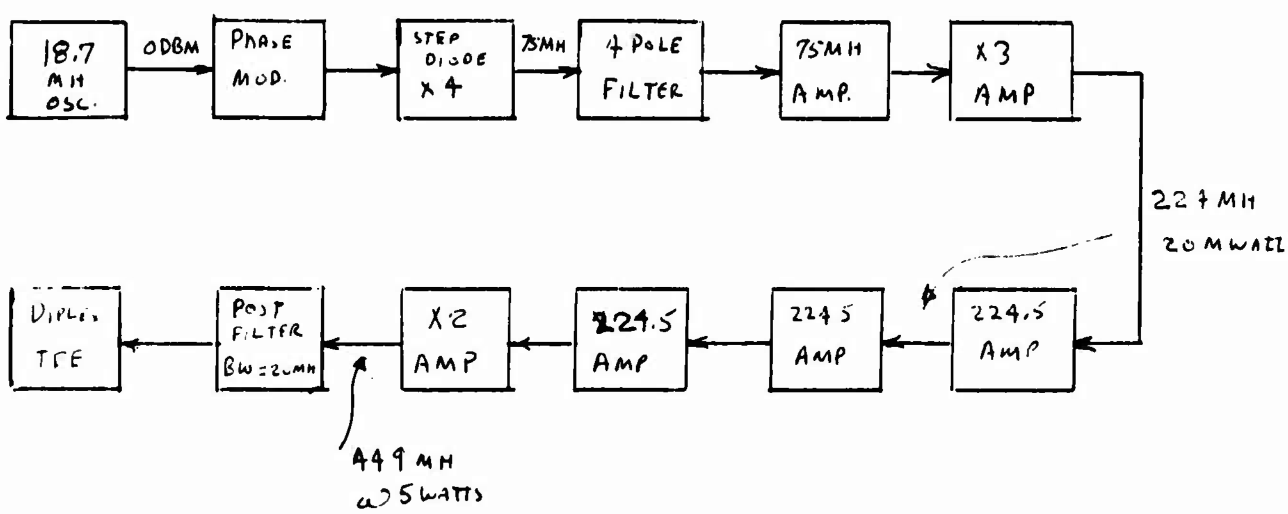
THE POST FILTER PROBLEM

At first, it became obvious that the choice of a post filter should be made on rejection of spurious (spec's MIL-11746) and minimum of poles to do this, so that output (post) filter would present a minimum loss to the 449 mc spectrum.

It then became obvious, from diplexing tests, that some of the 449 mc modulation spectrum was entering the receiver bandpass spectrum about 421 mc. Thus, the choice of the post filter should not be made on a spurious criteria, but on the amount of noise being generated by the transmitter, in the receiver acceptance band.



Case 1: Assume ideal noiseless transmitter (i.e., NF = 0 db).

$P_{g\,\text{Transmitter}}$ = Power gain of transmitter = $\dfrac{5 \text{ watts}}{1 \text{ milliwatt}}$ = 5000 or 37 db

K_T = -174 dbm/Hz

F = 0 db

B = 20 mc (i.e., determined primarily by the post and prefilters)

B = 73 db .

Amount of noise from transmitter generated by noise, KT only*, at preselector output (20 mc BW), assuming no post filter. (Assume no significant insertion loss in preselector for 421 mc $\pm$ 10 mc band)

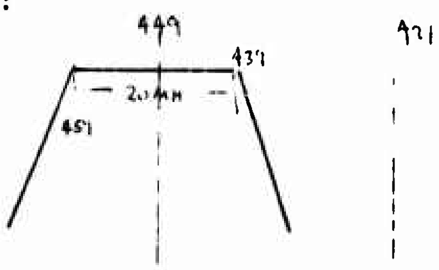
		-KT	B	F	Pg		
Amount of noise from transmitter at preselector output	=	- 174	+ 73	+ 0	+ 37	=	-64 dbm**
		KT	B	F			
Amount of noise at preselector output due to receiver without Diplexing transmitter	=	- 174	+ 73	+ 6		=	-95 dbm.

$\therefore$ minimum post filter attenuation required of transmitter output at 421 mc

= 95 - 64 = 31 db.

This is, of course, for our case of an ideal transmitter noise figure of 0 db.

For a bandpass post filter whose bandwidth = 20 mc and a 6 db/oct/pole slope, we have:



4 Pole

3 db at 439
27 db down at 429
51 db down at 404 mc

---

* Assuming Gaussian type noise is a fair approximation, even though class C stages are used.

** Assuming that all multipliers and amplifiers of transmitter are wideband (i.e., generate noise over BW, say $\gg$ +10 mc output BW related to 449 mc and $\approx$ constant amplitude). This is, of course, our case.

From the above, we would need a minimum 4 pole post filter to reject the transmitter noise, in the receiver band, to a level equal to that attributed to the receiver alone.

Case 2. Same as Case 1, except NF of transmitter = 10 db.

The result would be the same as in Case 1, except we would require a 41 db rejection at 421 mc.

4 pole filter would be a bare minimum

Case 3. Same as 2, except that we must realize that the transmitter noise has to be << that generated by the receiver front end alone, to ensure no appreciable receiver sensitivity degradation.

For a 10 percent degradation, this would mean that the noise from the transmitter should be 10 db down from that generated by the receiver (assuming uncorrelated noise sources).

The result is a minimum post filter attenuation of 51 db at 421 mc.

5 Pole

3 db down - 439 mc
33 db down - 429 mc
63 db down - 409 mc.

Minimum 5 pole filter.

Case 5. Actual measured case using 3 pole post filter of 20 mc bandwidth. Receiver sensitivity hurt by $\simeq$ 20 db when Diplexing.

Amount of noise power at receiver input due to receiver	=	-95 dbm
Amount of noise power at receiver input due to receiver and transmitter	=	- 95 + 20 -75 dbm
Amount of effective filtering by post filter at 421 mc	=	30 db calculated 26 db measured.

Amount of transmitter noise in 20 mc bw before post filter = -75 dbm+26 = -49 dbm.

What transmitter NF does this represent:

$N.F_{\text{Transmitter, calculated from measurements}} = -174 + 73 - 37 + NF = 49$ dbm

$NF_{\text{Transmitter}} = 15$ db .

Case 5. How many poles are required in the post filter to not degrade the NF of the receiver by more than 10 percent when the transmitter NF = 15 db?

Amount of rejection required of post filter at 421 mc $= -174 + 73 + 15 + 10 + 37 + 95$

$= 56$ db.

<u>6 Pole Minimum (Need 7)</u>

3 db - 439
39 db - 429
73 db - 409

Case 6. Same as Case 5, except instead of 10 percent, we take 1 percent of total receiver noise.

Amount of post filter rejection at 421 mc $= 66$ db

<u>7 Pole Minimum (Need 8)</u>

3 db - 439
45 db - 429
87 db - 409

Case 7. What is the NF of the last transmitter stage alone?

13 db loss in receiver sensitivity using only last stage as only wideband noise generation source (Sierra 420A is ≈ 1 mc BW and used as driver to last stage).

Amount of noise at preselector output due to receiver alone $= -95$ dbm

Amount of noise at preselector output due to transmitter last stage $= -95 + 13 = -82$ dbm.

Last transmitter stage, noise related to it (other side of A3-pole post filter)	=	-82 dbm + 26 dbm = -56 dbm
Last transmitter stage power gain	=	5 db
Last stage transmitter noise referred to input	=	-56 - 5 = -61 db

KTB noise at last stage transmitter inp[illegible] = -174 + 7 = 101 dbm

$\therefore$ last stage NF = 101 - 61 = <u>40 db</u>

Conclusions

1. Require minimum of 7 poles to satisfy filtering requirements of post filter.
2. NF of transmitter of 15 db is reasonable.
3. NF of last stage of 40 db is also reasonable.

APPENDIX B

SYSTEM BANDWIDTH REQUIREMENTS

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The spectral distribution of a signal, which has been frequency modulated by multiplier tones, has been treated by L. J. Gracoletts, in the Proceedings of the IRE, July 1947. Basically, the sideband distribution consists of all the possible combinations of the sum and differences of the Bessel functions of the individual modulating tones taken six (6) at a time. The equation of the distribution is:

$$a = A_0 \prod_{i=1}^{\infty} \left[\sum_{n_i=-\infty}^{\infty} J_{n_i}(z_i) \right] \genfrac{}{}{0pt}{}{\sin}{\cos} \left[\left\{ \omega_0 (1 + C_0) + \sum_{i=1}^{\infty} n_i \omega_i \right\} + \sum_{i=1}^{\infty} n_i \theta_i \right]$$

where Z is the mod index of signal i. The product of the summation will create all possible combinations of sums and differences of the various Bessel functions. The spectrum is derived for six (6) tones, each with a mod index of 2.4. 2.4 was chosen in order to reduce the number of terms, i.e., $J_0(2.4) = 0$. The actual signal format in MAT is six (6) tones, each with a mod index of 2.5.

To simplify, the problem was broken into blocks of combinations. For example, there are sixty-four (64) combinations created by f_1, f_2, f_3, f_4, f_5 and f_6 by the first order Bessel function product of each of the signal J_1^6. The number of combinations thus derived was 124,588. Following is a list of these combinations, their individual amplitudes, power in each sideband, the number of spectral lines of particular combinations and the total power contained in these combinations.

Frequency Product	Bessel Functions	π Amplitude	$(\pi)^2$ $(\text{Amplitude})^2$	Σ No. of Spectral Lines	Total Power
$f_1, f_2, f_3, f_4, f_5, f_6$	J_1^6	.01982	.00039828	64	.025140
$2f_1, f_2, f_3, f_4, f_5, f_6$	J_1^5, J_2	.0164211	.0002696	384	.103500
$2f_1, 2f_2, f_3, f_4, f_5, f_6$	J_1^4, J_2	.013605	.0001851	960	.177700
$2f_1, 2f_2, 2f_3, f_4, f_5, f_6$	J_1^3, J_2^3	.01127245	.000127068	1280	.162600
$2f_1, 2f_2, 2f_3, 2f_4, f_5, f_6$	J_1^2, J_2^4	.0093384	.0000872057	960	.083717
$2f_1, 2f_2, 2f_3, 2f_4, 2f_5, f_6$	J_1, J_2^5	.007735	.00005983	384	.022970
$2f_1, 2f_2, 2f_3, 2f_4, 2f_5, 2f_6$	J_2^6	.006409	.00004421	64	.002829
$3f_1, f_2, f_3, f_4, f_5, f_6$	J^5, J_3	.007544	.00005691	384	.021850
$3f_1, 2f_2, f_3, f_4, f_5, f_6$	J^4, J_2, J_3	.006251	.00003907	1920	.075010
$3f_1, 2f_2, 2f_3, f_4, f_5, f_6$	J_1^3, J_2^2, J_3	.005179	.000026822	3840	.102996
$3f_1, 2f_2, 2f_3, 2f_4, f_5, f_6$	J_1^2, J_2^3, J_3	.004289518	.00001840	3840	.070660
$3f_1, 2f_2, 2f_3, 2f_4, 2f_5, f_6$	J_1, J_2^4, J_3	.003554	.00001263	1920	.024250
$3f_1, 2f_2, 2f_3, 2f_4, 2f_5, 2f_6$	J_2^5, J_3	.002944	.000008667	384	.003328
$3f_1, 3f_2, f_3, f_4, f_5, f_6$	J_1^4, J_3^2	.001826	.000003334	960	.003201
$3f_1, 3f_2, 2f_3, f_4, f_5, f_6$	J_1^3, J_2, J_3^2	.002379	.000005660	3840	.021730
$3f_1, 3f_2, 2f_3, 2f_4, f_5, f6$	J_1^2, J_2^2, J_3^2	.001971	.000003885	5760	.022380

Frequency Product	Bessel Function	π Amplitude	$(\pi)^2$ (Amplitude)^2	Σ No. of Spectral Lines	Total Power
$3f_1, 3f_2, 2f_3, 2f_4, 2f_5, f_6$	J_1, J_2^3, J_3^2	.001633	.000002667	3840	.010240
$3f_1, 3f_2, 2f_3, 2f_4, 2f_5, 2f_6$	J_2^4, J_3^2	.001353	.000001831	960	.001758
$3f_1, 3f_2, 3f_3, f_4, f_5, f_6$	J_1^3, J_3^3	.001093	.000001195	1280	.001529
$3f_1, 3f_2, 3f_3, 2f_4, f_5, f_6$	J_1^2, J_2, J_3^3	.000905	.000000819	3840	.003145
$3f_1, 3f_2, 3f_3, 2f_4, 2f_5, f_6$	J_1, J_2^2, J_3^3	.000750	.0000006006	3840	.002306
$3f_1, 3f_2, 3f_3, 2f_4, 2f_5, 2f_6$	J_2^3, J_3	.0006214	.0000003867	1280	.000495
$4f_1, f_2, f_3, f_4, f_5, f_6$	J_4, J_1^5	.002450	.000006003	384	.002305
$4f_1, 2f_2, f_3, f_4, f_5, f_6$	J_4, J_2, J_1^4	.001987	.000003948	1920	.007580
$4f_1, 2f_2, 2f_3, f_4, f_5, f_6$	J_4, J_2^3, J_1^2	.001182	.000002820	3840	.010800
$4f_1, 2f_2, 2f_3, 2f_4, f_5, f_6$	J_4, J_2^3, J_1^2	.001393	.000001940	3840	.007450
$4f_1, 2f_2, 2f_3, 2f_4, 2f_5, f_6$	J_4, J_2^4, J_1^1	.001154	.000001331	1920	.002556
$4f_1, 2f_2, 2f_3, 2f_4, 2f_5, 2f_6$	J_4, J_2^5	.00095614	.0000009142	384	.00351
$4f_1, 3f_2, f_3, f_4, f_5, f_6$	J_4, J_3^1, J_1^4	.000932	.0000008686	1200	.001040
$4f_1, 3f_2, 2f_3, f_4, f_5, f_6$	J_4, J_3, J_2, J_1^3	.0007726	.0000005969	7680	.004584
$4f_1, 3f_2, 2f_3, 2f_4, f_5, f_6$	J_4, J_3, J_2^2, J_1^2	.0006400	.0000004096	23040. 987093	.009440
$4f_1, 3f_2, 2f_3, 2f_4, 2f_5, f_6$	J_4, J_3, J_2^3, J_1	.0005301	.0000002810	7680	.002158
$4f_1, 3f_2, 2f_3, 2f_4, 2f_5, 2f_6$	J_4, J_3, J_2^4	.00043936	.00000019304	1920	.000371
$4f_1, 3f_2, 3f_3, f_4, f_5, f_6$	J_4, J_3^2, J_1^3	.000355	.000000126	3840	.000484
$4f_1, 3f_2, 3f_3, 2f_4, f_5, f_6$	J_4, J_3^2, J_2, J_1^2	.0002940	.000000086436	11520	.000996
$4f_1, 3f_2, 3f_3, 2f_4, 2f_5, f_6$	J_4, J_3^2, J_2^2, J_1	.000243568	.000000059325	11520	.000683
$3f_1, 3f_2, 3f_3, 3f_4, 2f_5, 2f_6$	J_3^4, J_2, J_1	.000344605	.000000118753	1920. 994360	.000228

The distribution of the different sums and differences were then derived and the power content in each megacycle at bandwidth was determined.

The spectral distribution is almost continuous and can only be graphically represented by its peak values and power density. Figure 1 is a graph showing the peak value of the spectral lines versus BW. The abscissa is in voltage compared to the unmodulated carrier value. It can be noted that 1 percent of maximum occurs at approximately 8.0 mcs on a two-sided bandwidth of 16.0 mcs. It should be remembered that the curve is a peak envelope, and the number of spectral lines would make the spectral distribution appear almost continuous. A more meaningful curve may be the power density distribution. Figure 2 shows the power distribution in density or power contained in each megacycle of two side bandwidth. We can note that the power density has dropped to 40 db below the total signal content at approximately 16.0 mcs. Figure 3A is a curve of the integrated power content as a function of bandwidth. As noted, 99 percent of the power is contained in 12 MHz of bandwidth.

Figure 3A represents the amount of total power contained within a frequency band (double sided), about center frequency, for six tones of mod index 2.4 each. This curve was derived without taking into consideration a filter amplitude response or assumed one in which the band of interest was flat and the skirts infinitely steep (i.e., a square response). This should not imply that since 99 percent of the power is contained in 12 MHz of bandwidth, then a filter 3 db bandwidth of 12 MHz can be used.

It can be seen from Figure 3B that this is far from true for practical filter responses, and especially those obtained in the MATS transponder.

Case 1. Filter response between 3 db points in Figure . A, BW_{3db} = 12 mc.

Diplexer Multipole Simulation

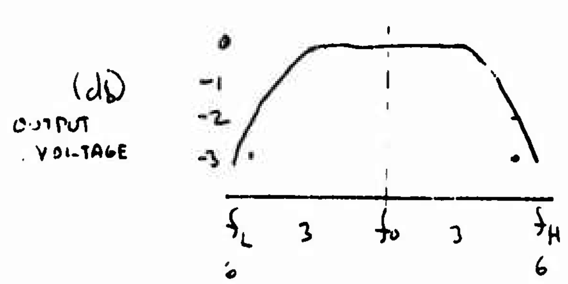


Figure A

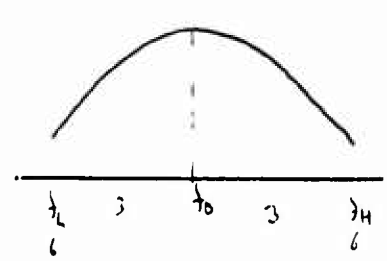
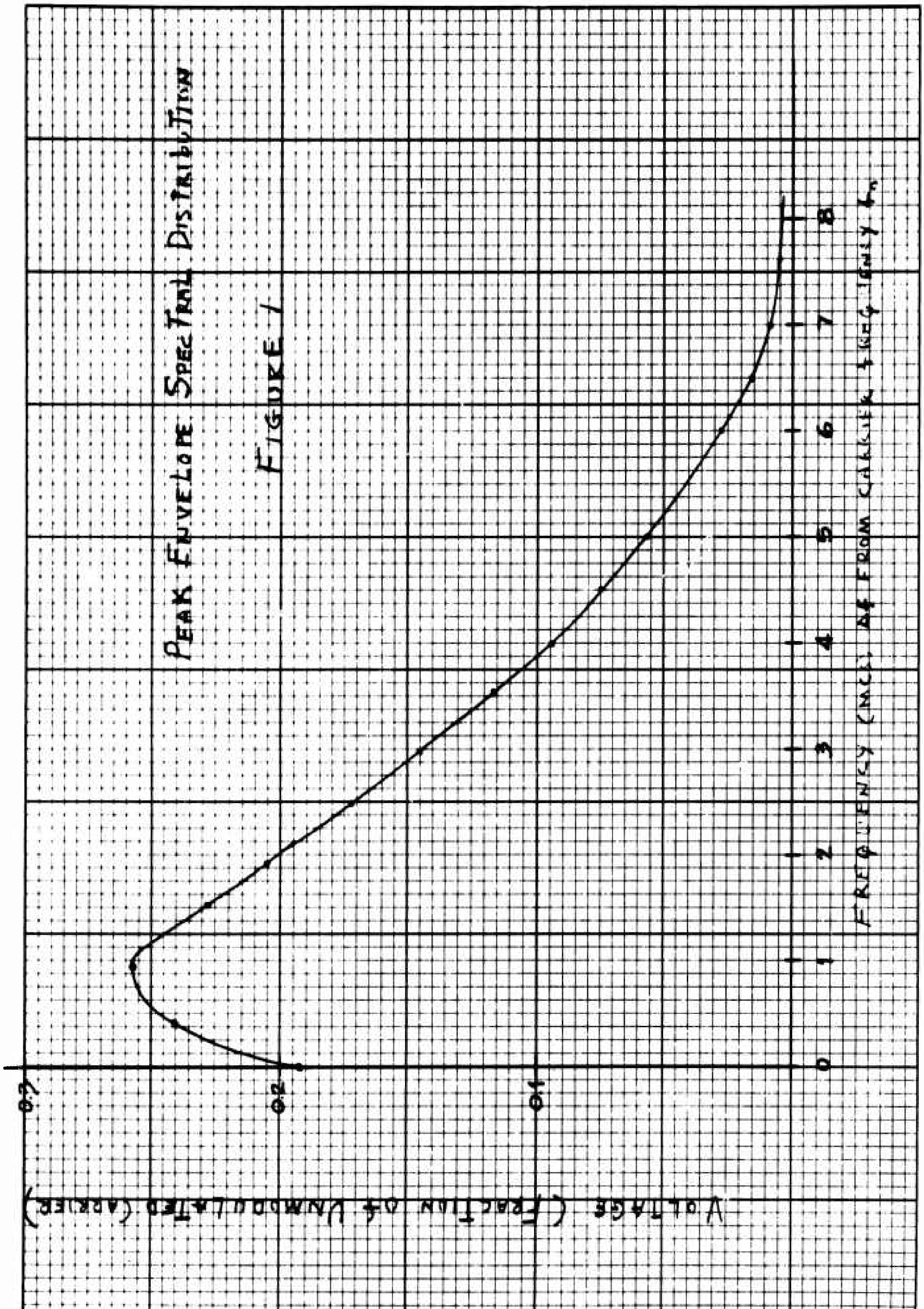
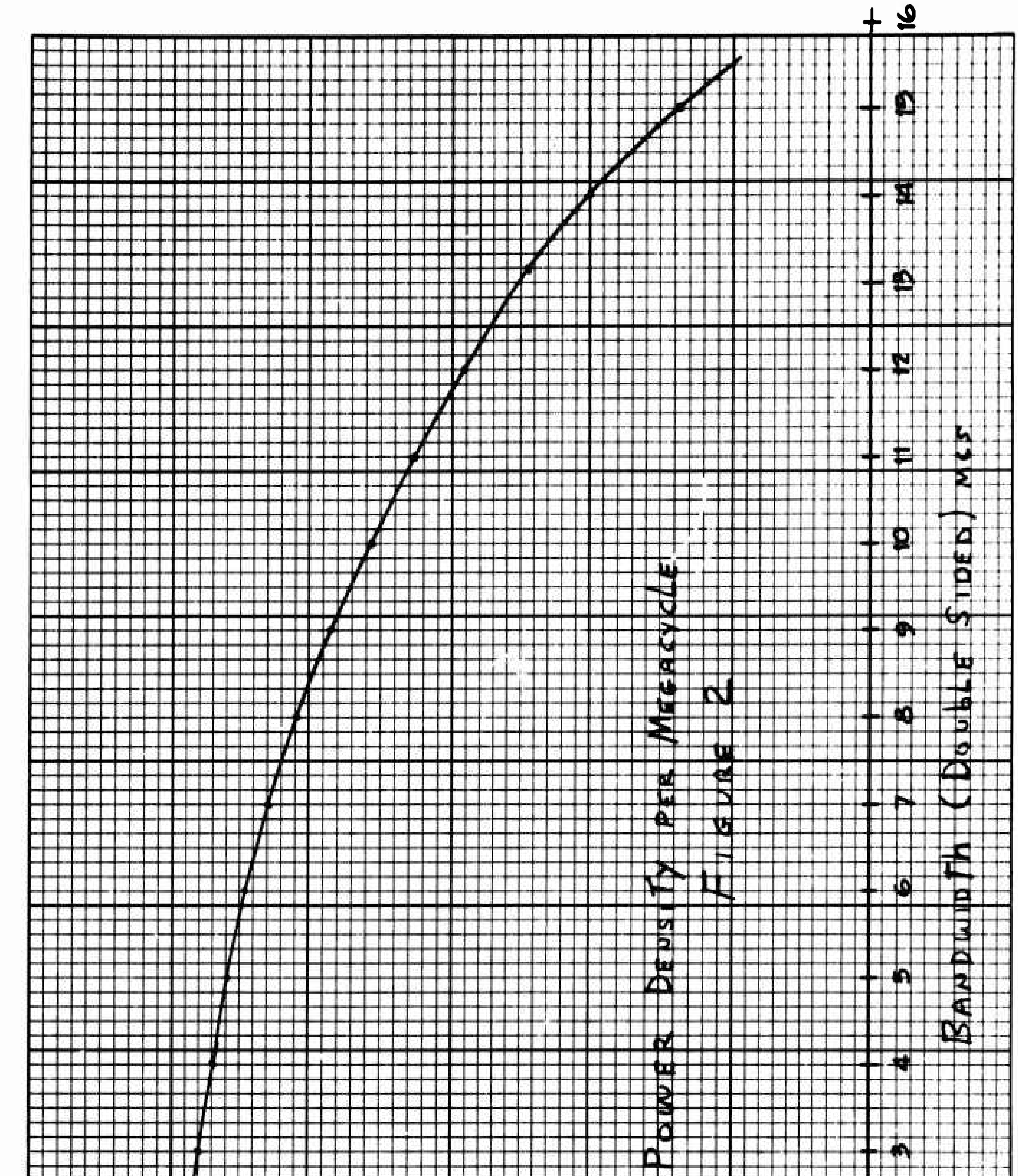
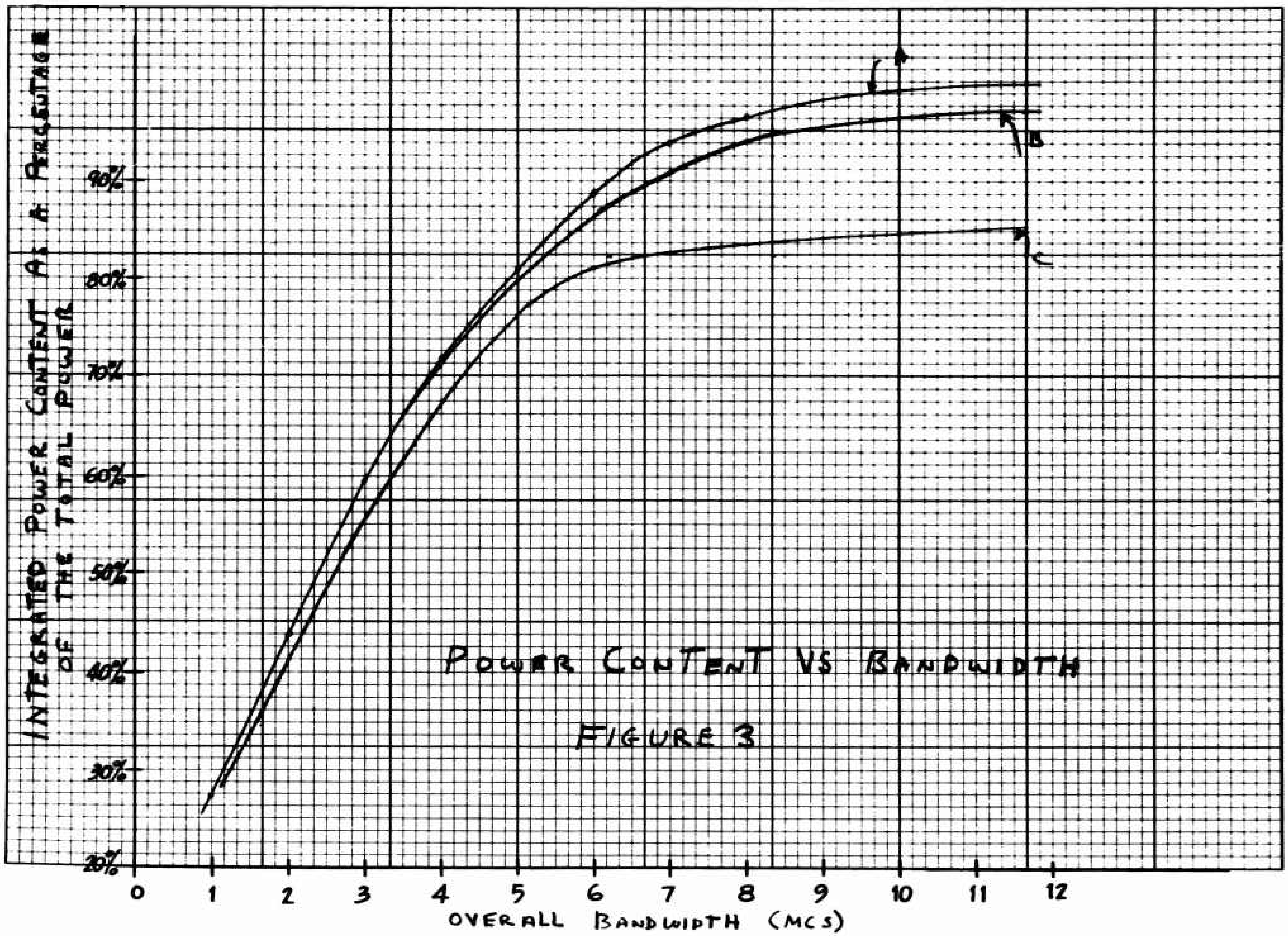


Figure B


PEAK ENVELOPE SPECTRAL DISTRIBUTION
FIGURE 1
0.3
0.2
0.1
VOLTAGE (FRACTION OF UNMODULATED CARRIER)
0
1
2
3
4
5
6
7
8
FREQUENCY (MC), Δf FROM CARRIER FREQUENCY fm


POWER DENSITY PER MEGACYCLE
FIGURE 2
BANDWIDTH (DOUBLE SIDED) MCS
POWER DENSITY PER MEGACYCLE DB REF. TO PEAK
0
-10
-20
-30
-40
-50
0
1
2
3
4
5
6
7
8
9
10
11
12
13
14
15
16



POWER CONTENT VS BANDWIDTH

FIGURE 3

Case 2. Single pole response due to transmitter output into post filter (assuming diplexer filter ideal square response), see Figure B and Figure 3, curve B, for 12 MHz bandwidth.

Thus, the bandwidth of the transmitter final amp and diplexer should be wider than 12 MHz in order to pass near 99 percent of the transmitter power.

APPENDIX C

NOISE BANDWIDTH CORRECTION FACTOR

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NOISE BANDWIDTH CORRECTION FACTOR

Since it is well known that single pole filters have 3 db bandwidths that are narrower than their actual noise bandwidths, then, when calculating S/N ratios for a receiver in which this type of filter is used a common correction factor is used.

$$\text{Noise Bandwidth} = \pi/2 \times (3 \text{ db Bandwidth}) \quad .$$

This factor is almost, but not quite, correct for the crystal filters used in the MATS transponder. The above $\pi/2$ factor is related to a single pole filter with a 6 db/oct skirt selectivity.

Since the MATS crystal filters are used in a PFFB loop, the slope is $\not<$ 6 db/oct. In addition, the common spurious crystal modes add to the over-all noise bandwidths.

Figure 1 is a closed loop voltage plot of a typical crystal. The 3 db bandwidth measured is 4.1 KHz. Figures 2 and 3 are plots of $(\text{voltage})^2$ versus frequency for the same filter. Calculating the noise power bandwidth graphically, we have

$$\text{Noise power bandwidth} = 9 \text{ KHz.}$$

$$Q = \frac{\text{Noise power bandwidth}}{\text{3 db bandwidth}} = \frac{9 \text{ kc}}{4.1 \text{ kc}} = 2.2, \text{ or } 3.4 \text{ db} \quad .$$

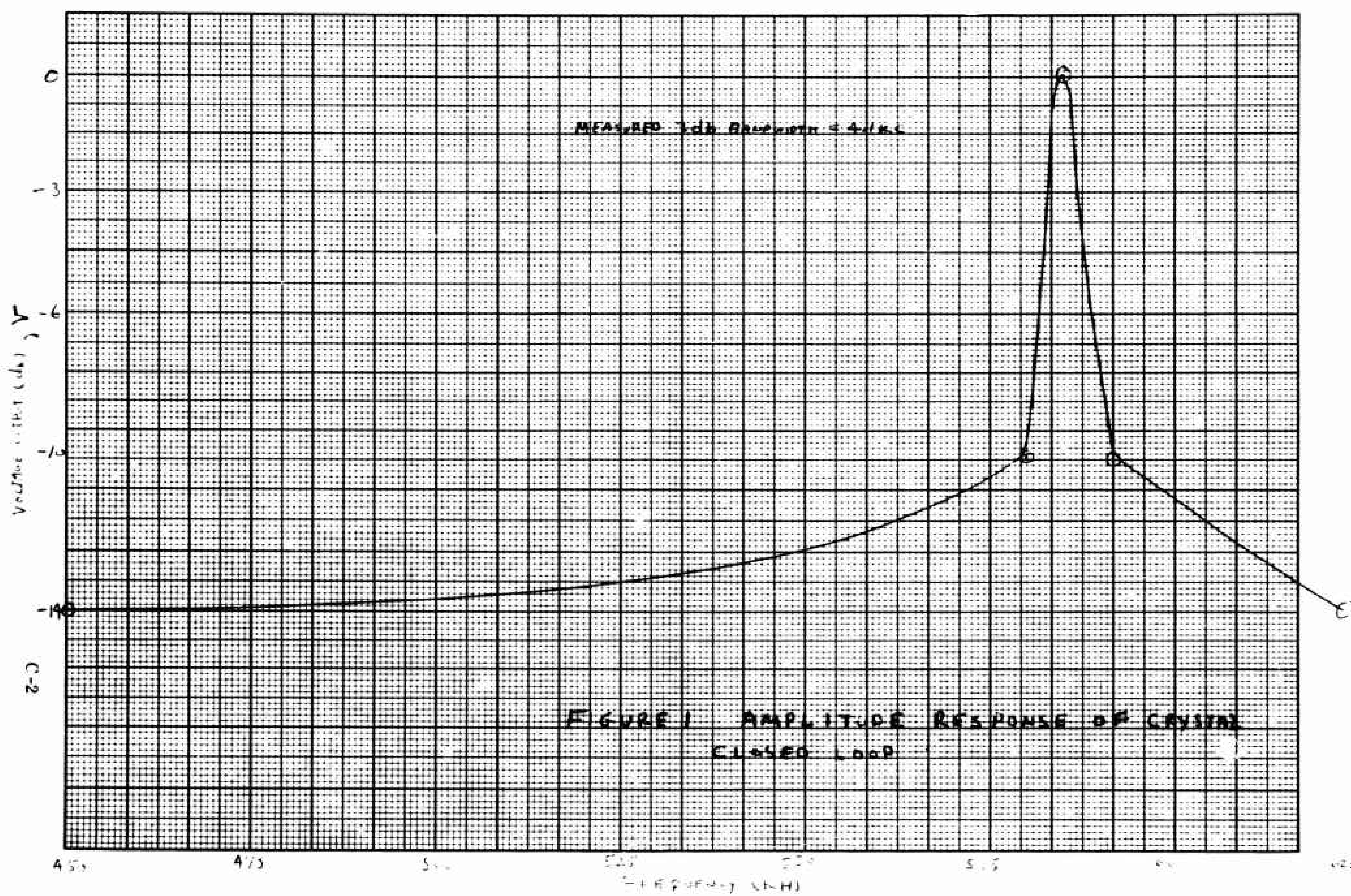

FIGURE 1 AMPLITUDE RESPONSE OF CRYSTAL
CLOSED LOOP

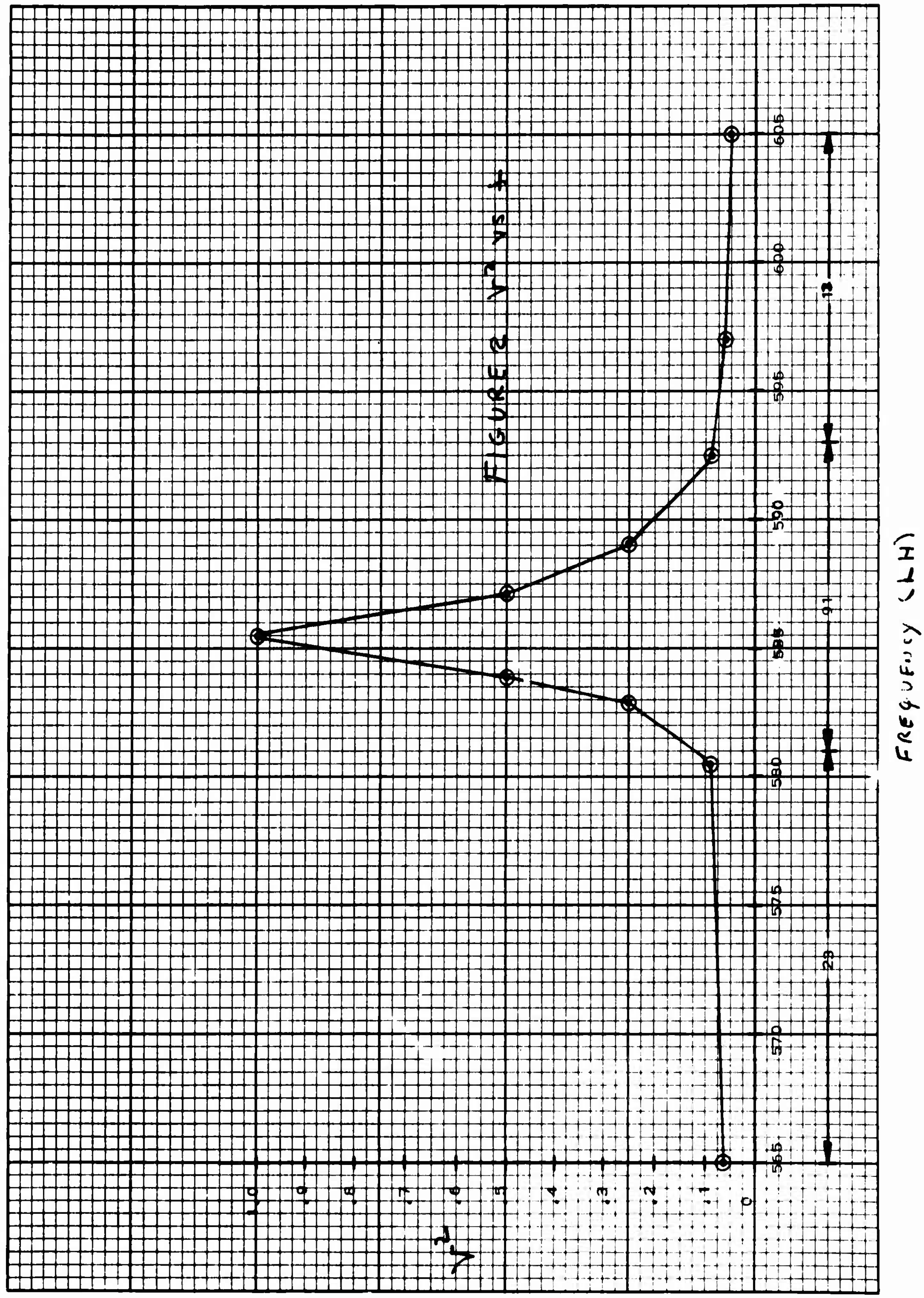
FIGURE 2 Y^2 vs f
Y^2
1.0
.9
.8
.7
.6
.5
.4
.3
.2
.1
0
565
570
575
580
585
590
595
600
605
23
16
13
FREQUENCY (LH)

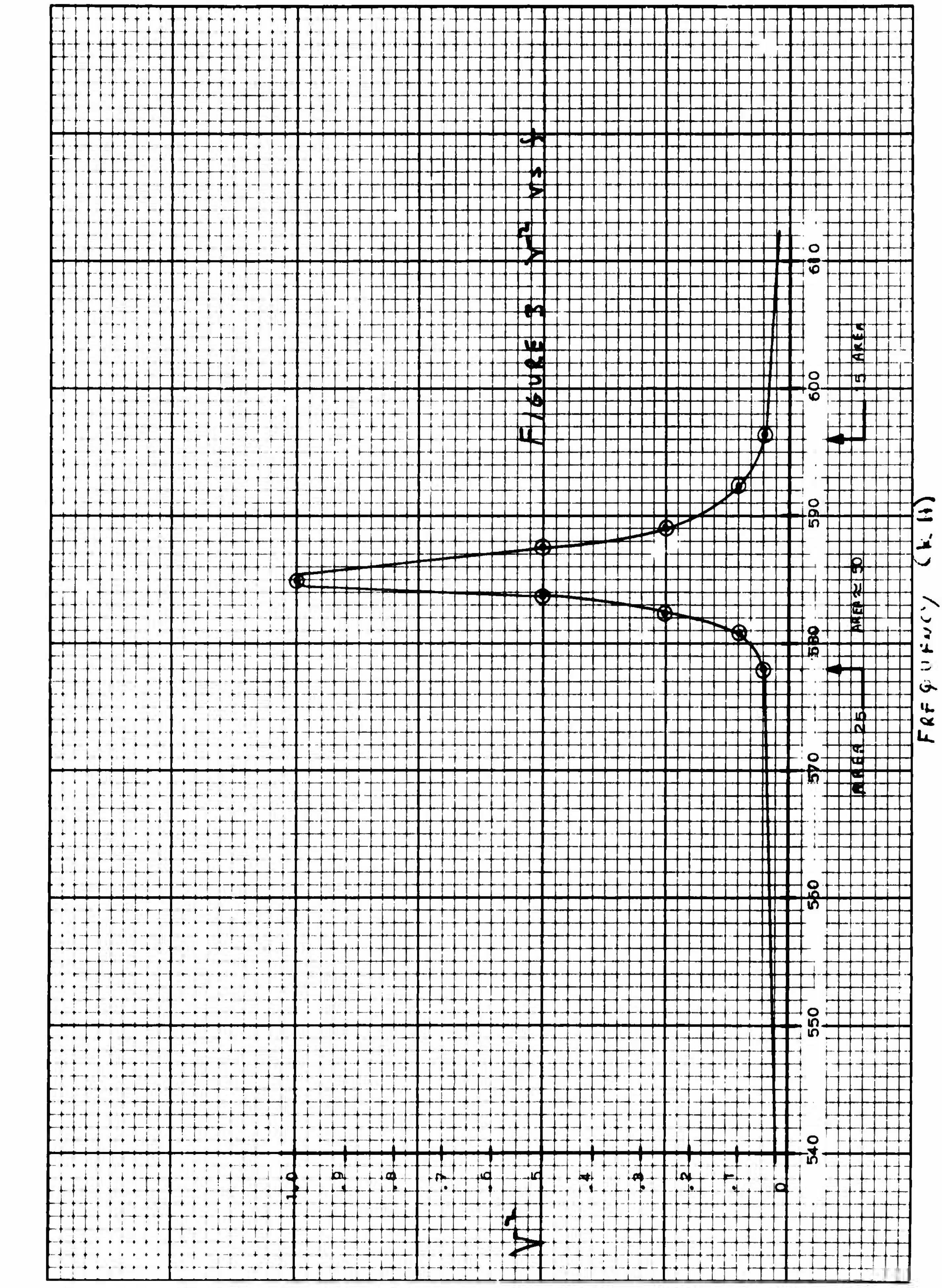
FIGURE 3 V² VS f
V²
1.0
.9
.8
.7
.6
.5
.4
.3
.2
.1
0
540
550
560
570
580
590
600
610
AREA 25
AREA ≈ 50
5 AREA
FREQUENCY (KH)